

Hooray for LC Filters!

—simplest is sometimes best

Design a simple filter and a useful control circuit to go with it.

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With names like Chebyshev, Butterworth, M-Derived, Maximally Flat, or Minimum Ripple, it is sometimes difficult to decide just how to pick a filter, particularly one that will pass or reject a small band of low frequencies. Well, why not use an LC circuit, since it is quite stable and simple to make?

The difficulty with the tuned LC circuit, of course, is that it is usually too sharp to pass or stop a band of frequencies, and has to be loaded (resistively) to reduce the Q. Fig. 1 shows the series (a) and parallel (b) forms of this circuit, along with a generalized impedance curve (c),

which shows how the impedance of the circuit changes with frequency and Q.

Another problem with this network is that as the notch or center frequency of the band of frequencies we would like to pass or stop is lowered, the necessary values of L and C become very large. Just calculate, for a center frequency of 160 Hz, the values of L and C for a series-tuned LC circuit using the following formula: $f = \frac{1}{2\pi\sqrt{LC}}$, where f = frequency in Hz, L = inductance in Henrys, and C = capacitance in farads. You will see why the parts can't be nicely fitted on a printed circuit board. In this age of miniaturization, it is difficult to even see a resistor, capacitor, or amplifier!

Don't despair—a circuit known as the Twin-T Filter

will not only fit on the board, but is extremely simple to design. The circuit layout is shown in Fig. 2.

All you have to know in order to design this network to "stop" or at least attenuate a particular frequency is the following set of relations: $C1 = 2C2$, $R1 = 2R2$, and $f = \frac{1}{2\pi R1C2}$. For example, suppose we want to design the filter to have a center frequency of 160 Hz. By rewriting $f = \frac{1}{2\pi R1C2}$ as $R1C2 = \frac{1}{2\pi f}$, we find $R1C2 = \frac{1}{6.28(160)} = 0.000995$. Since there are an infinite number of ways to multiply $R1$ and $C2$ in order to get the value 0.000995, we have to narrow things down a bit. A simple way to do this is to pick a large value for C that is available in a small physical size, such as a high K (dielectric constant) ceramic-type capacitor. Let's try .01 microfarads ($.01 \times 10^{-6}$ farads). Using this value, we find that $R1 = 9.95 \times 10^4 / .01 \times 10^{-6} = 9.95 \times 10^2 / .01 = 9.95 \times 10^4 \approx 100k$ Ohms.

Therefore, with $C2 = .01$ uF, then $C1 = 2C2 = .02$ uF and $R2 = R1/2 = 50k$ Ohms. Using these values, with small ceramic capaci-

tors and 1/4-Watt resistors, we can produce a filter that will pass all frequencies except those directly around the center frequency of 160 Hz. This is a notch filter.

Let's assume that we want to stop all frequencies, but we would like to pass the frequency of 160 Hz. The op amp circuit outlined in Fig. 3 does this with good stability.

This is a negative feedback circuit, which means that the output of the amplifier is added to the input such that the input to the amplifier is reduced. The amount of feedback is controlled by the feedback network, which in our case is the Twin-T filter. Since this network passes all frequencies except a small band of frequencies around the design center (160 Hz), the feedback is large at all these frequencies, and the input to the amplifier will, therefore, be very small. This means that the output will be very low. In the region of 160 Hz the feedback is almost zero, so the output becomes large.

In other words, if we attach an audio oscillator to the input of this circuit, and slowly increase the frequency starting at about 10

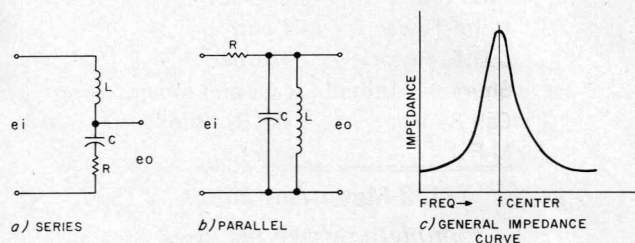
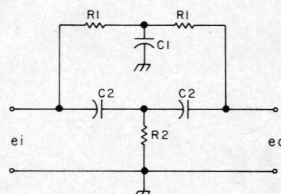


Fig. 1. Series (a) and parallel (b) tuned circuits with generalized impedance curve (c).

Hz, we will get no output until the dial reads from about 130 to about 190 Hz, with the signal peaking at 160 Hz. As the frequency is increased further, the output will again drop to zero.



2. Twin-T filter schematic diagram.

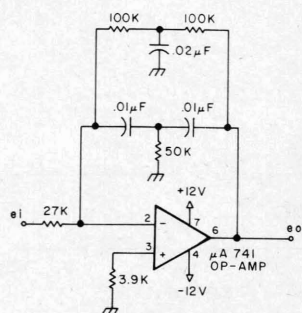


Fig. 3. 160-Hz bandpass filter schematic diagram.

This is similar to the general tuning curve of the LC circuit shown in Fig. 1(c).

This circuit was used in a tone-operated slide viewer that a friend asked me to build for him. The tape recorder that he used to

provide the narrative for each slide, as it was shown by the projector, contained a tone of 160 Hz which occurred at the end of each narration. A microphone near the speaker picked up the tone, which was then amplified, and this oper-

ated a relay which caused the slide mechanism to insert a new slide. This circuit is shown in Fig. 4, and it should be self-explanatory. With a little ingenuity, many uses can be devised for this simple filter. ■

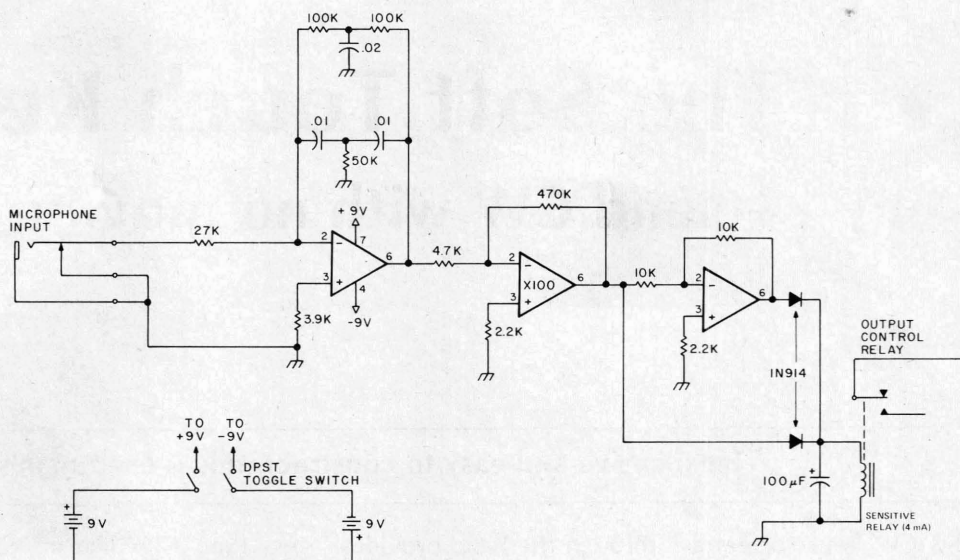


Fig. 4. Automatic slide-advance system using a 160-Hz Twin-T filter for cueing from a tape recorder speaker.

DX

from page 55

and unless some decision is made soon by the Iraqi officials, that may be the end result.

HB9MX reports that he hit bedrock in the mountain of Clipperton QSLs and is now answering QSLs as they come in. As of the end of September, he had answered 19,298 QSLs. 15,812 were returned direct by air mail, with the remainder going via the various bureaus. 852 cards were rejected because they were not in the log. If you still don't have your Clipperton QSL, you can now get an immediate response from Kurt Bindschedler, Strahleggweg 28, 8400 Winterthur, Switzerland.

Australian Novices, recognizable by the letter N immediately following their district number in their call sign, are allowed to run ten Watts on CW and thirty Watts on SSB. They operate in the following subbands: 28100-28550, 21125-21150, and 3525-3625 on CW, and 28100-28600, 21150-21200, and 3525-3625 on SSB.

9V1TT is apparently the only XYL active in Singapore. She is the XYL of 9V1TP.

Ownership of Abu Ail, once considered unclaimed, seems to be a bone of contention between France and Djibouti. If it is owned by Djibouti, then it is just another offshore island. But if Abu Ail belongs to France, then it would be a separate DXCC country.

FH8CY should become a bit more available now that the Northern California DX Foundation has donated an FT-101 and outboard vfo. This was the rig used by K5YY on his recent African swing.

The efforts of the Northern California DX Foundation have put many new ones in the pockets of the deserving DXers. Drop them a line at PO Box 717, Oakland CA 94604, to see how you can get in on the fun.

DJ9ZB is publishing a large, plastic-bound QSL book which lists a large number of QSL Managers. It contains some 50 pages of listings, and is a great aid for the serious DXer. Order from DJ9ZB, Carl-Kistner Str. 19, D-7800 Freiburg, West Germany.

If you still need cards from the SV0WZ stay on Rhodes, try WB3JRL.

Well, that about covers the DX front for this month. We want to again thank the *West Coast DX Bulletin*, the *LIDX*

Bulletin, and *WorldRadio News* for much of the preceding information. We also want to thank all you readers who have sent reports and photos. Keep them coming, and we will see you again next month.

Ham Help

This letter is in regard to the article "Rejuvenate A Pawnee," in the October, 1978, issue. I liked the way K4GRT and W4IEV laid out the modifications and diagrams—even I could follow them. I would like to know if someone else will do the same sort of article on the Hallicrafters SR-42 2 meter AM

rig, as that is what I have and I would like to put it on 2 meter FM.

If an article is not possible, is it possible that someone would modify it for me at a reasonable price? Thanks.

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Corrections

Please note the following corrections to my article "The Klassic Kilowatt," which appeared in the November, 1978, issue: On page 226, column 2, line 9, should read: "3 dB weaker than a large." On page 228, column 2, line 17 should read: "output power is 700 Watts." (It is assumed that the reader realizes that the amplifier is loaded into a dum-

my load for these measurements.) On page 229, column 1, lines 12 through 14, should read: "are 1500 volts at 700 mA on 20 meters, producing 700 Watts output." (This power was measured using a Drake W4 wattmeter while feeding a nonradiating dummy load.)

Dave Ingram K4TJW
Birmingham AL

Explore the World of VLF

— with this simple converter

A broadband preamp makes this unit tick.

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Woodbine MD 21797

After looking at available VLF converter data, I embarked on a design to facilitate some

VLF listening. Because it was convenient relative to the communications receiver that I planned to use the converter with, a 10 MHz i-f was picked. This removed the i-f far from many possible sources of i-f feedthrough.

The circuit shown utilizes an LM318 as a

broadband op amp, providing greater than 8 dB gain from 20 kHz up to 400 kHz. The mixer utilized is a 40673 MOSFET containing internal protective diodes (not indicated in the schematic). See Fig. 1. The local oscillator (LO) employs a 2N5458 FET as a rather conventional crystal oscillator operating at 10 MHz. LO injection is controlled by C_{inj} and tuning by C_t , although, of course, there is some interaction between the two capacitors.

I used a low-frequency signal generator with calibrated output, and I adjusted C_{inj} for best sensitivity with a very weak signal generator output, while maintaining the LO frequency at 10 MHz. The receiver used had a tuning range from 10 to 10.6 MHz, yielding a converter tuning range from zero beat to 600 kHz. Other appropriate frequencies could be employed to accommodate the particular receiver

to be used. It is recommended that the LO frequency be kept relatively high, say, at at least 40 meters, to ameliorate i-f spurious responses as previously mentioned. If a 40 meter ham band receiver is to be employed, simply use a 7.0 MHz crystal and retune the LC tank in the drain circuit of the 2N5458 FET. It may be necessary to add a bit of fixed capacitance in parallel with the existing 3-30 picofarad trimmer in order to tune down to 7.0 MHz.

The lower frequency limit of usability is, at present, about 10 kHz because of the 10 MHz LO feedthrough into the receiver. I think that a new layout would have to be made to further preclude this problem. A two-section low-pass filter is utilized on the input to the LM318 to preclude cross-modulation with strong signals in the broadcast band and above. The filter rolls off 6 dB by 500 kHz and is 20 dB down

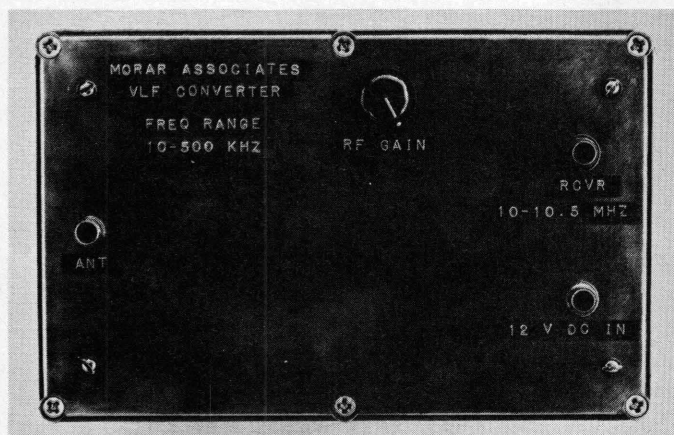


Photo A. Front panel, showing antenna input on left, 10 MHz out in the upper right, and 12 V dc input in the lower right. A tunable i-f output was added after this picture was taken (see the schematic). The tunable inductance core shaft was brought out as an operating control slightly to the right of the rf gain control. It can be seen in Photo C.

at 750 kHz.

The converter was tried with several antenna arrangements. The most satisfactory was a 1000-foot longwire antenna that terminates in the shack. It still left much to be desired, so a separate antenna coupler was assembled. See Photo C. This greatly improved the reduction of cross-modulation and converter-usable sensitivity. Even with the low-pass filter at the converter input, strong local stations in the broadcast band plus local 100 kHz loran signals would cause objectionable cross-modulation problems without the antenna coupler.

I am also working on a fixed tuned TRF receiver for receiving WWVB, wherein a double-shielded tuned rotatable 27"-radius loop will be employed.

With the antenna coupler indicated in Photo C, it was possible to receive WWVB at 60 kHz at this location with an S-meter reading of S9, indicating an input in the vicinity of 20 microvolts. Without the antenna coupler, it was not possible to hear WWVB at all due to spurious responses about either side of 60 kHz. The converter requires 12 volts at 12 mA.

As can be seen from the accompanying photos, the converter is built on a 3½" x 6" single-clad phenolic board, utilizing vector mini-clip pins pushed into the holes of isolated copper pads, produced by an isolated-pad-Drill-Mill tool. See Photo B. The converter is housed in a Bud CU-247 Econo-box, with dimensions of 7-25/64" x 4-45/64" x 2-7/32". The antenna coupler is in a separate box, since it came as an afterthought and would have made the dimensions and layout of the converter unduly large anyway.

Fig. 1. shows the antenna coupler converter and Ken-

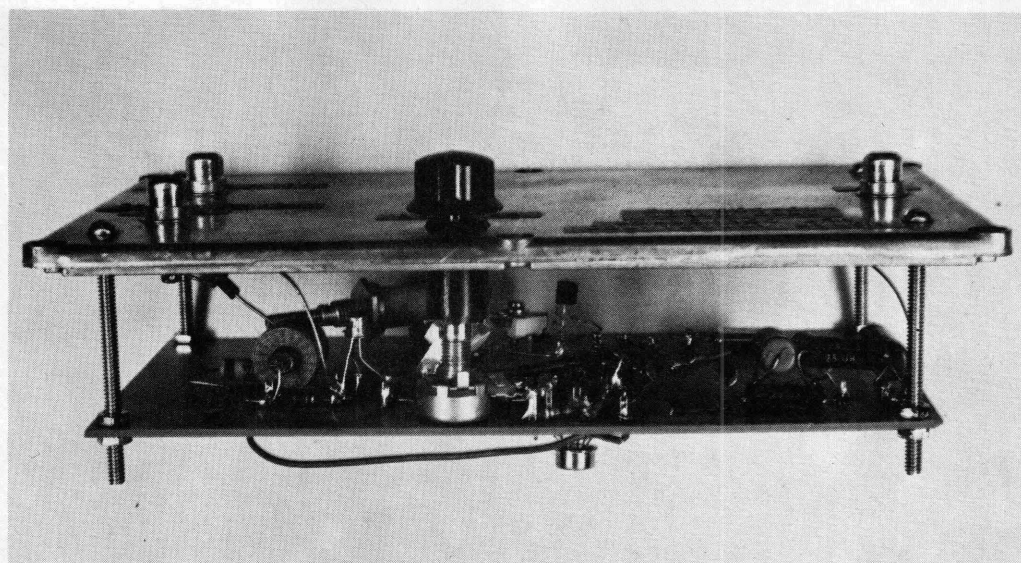


Photo B. Side (open view) of the converter showing the copperclad board and vector mini-clip construction described in the text.

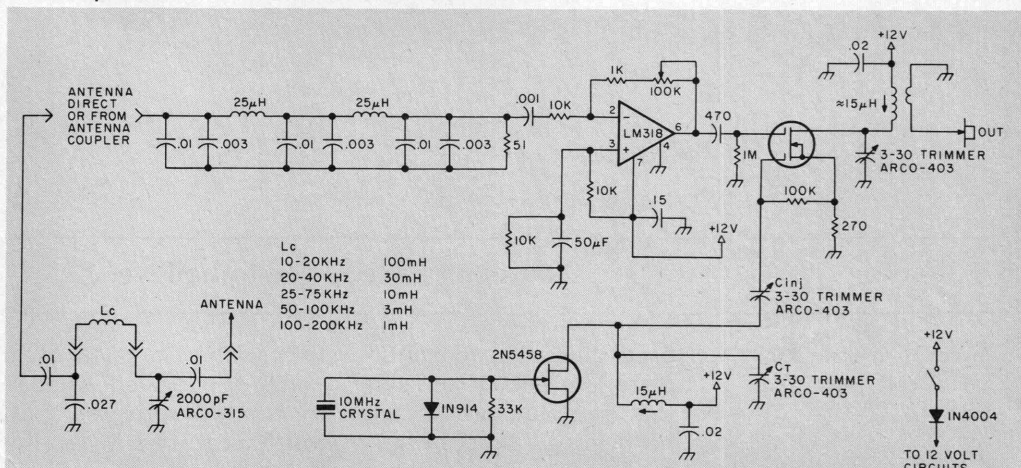


Fig. 1. VLF converter.

wood R599 receiver set up and connected to the 1000-foot longwire antenna which is 60 feet high. The antenna coupler was built in a 3" x 5" x 7" minibox, Bud CU3008. See Photo C. Coils were mounted on the selector switch so as to be as mutually perpendicular to each other as possible to minimize magnetic coupling. The Arco-type 315 compression capacitor has a range of 2525 pF tight and 1200 pF at 3 turns open. Using the inductances indicated, satisfactory tuning was obtained from 20 kHz to 200 kHz. Above 200 kHz, the converter had plenty of gain connected directly to the antenna without the coupler. Cross-modulation was not a problem in this region, either. ■



Photo C. The converter set up with the Kenwood R599 receiver. The antenna coupler can be seen at the top above the converter.

Build This Excitingly Simple Receiver

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You may or may not be familiar with the various versions of miniaturized communications receivers I've presented over the past 6 years as part of my Minicom series. Although I'm up to an MK VI version at this writing, circumstances have arisen which prompted a decision to revisit the MK III.

The Original MK III

The original MK III Minicom¹ was an electronically tuned 80 meter receiver designed around the Motorola MVAM1 triple varactor. This solid state device took the place of the normal 3-gang variable capacitor used to tune the rf, mixer, and vfo tank circuits. At \$13.50 for the MVAM1, I considered the

receiver to be more of a novelty than anything else. I included it in my article merely to show existing possibilities to further shrink receiver size, never dreaming of the kind of interest it would arouse.

As has been said about most living things, they start to die the moment they're born. The MK III may not have been a living thing, but it sure died in a hurry. While the article was awaiting publication, Motorola ceased production of the MVAM1. Although the MK III was only one of several ideas covered in the original article, almost everyone who wrote to me was interested solely in duplicating the MK III and nothing else. The question of whether or not they'd have gone ahead after learning what it would cost became academic at that point, and all I could do was pass along the bad news and watch the

tears.

New Hope

After the demise of the MVAM1, Motorola introduced the MVAM115 single varactor diode. It, too, was designed for electronic tuning of AM radios and had a high capacitance ratio along with a Q of 150 or better. To make it even more worthwhile, 3 of the new diodes would cost only about a third the price of the old MVAM1. With all this going for it, it was time to bring the MK III back to life. In short order, I obtained some MVAM115s and got to work.

The results were all I could have hoped for during the breadboard experiments, and, as soon as all the variables had been optimized, a PC was laid out and a finished receiver built. Surprisingly, it worked right off, and all I could assume was that Murphy was on vacation or

had passed on.

Why Diode Tuning?

Besides a considerable saving in bulk, electronic tuning raises some other interesting possibilities. The tuning pot is connected by wire to the receiver assembly and thus allows remote tuning, if desired. The PC portion of a receiver can be tucked away in a corner of the cabinet, while the tuning pot can be panel mounted without the worry of mechanical linkage to turn the shaft of a variable capacitor. Preset voltages can be switch selected for often used frequencies. A second pot controlling a fraction of a volt can have its output summed with the main tuning voltage to allow band-spreading over very small portions of the dial. In the modern vernacular, this is known as "incremental tuning." Other ideas, such as being able to reverse the voltage so as to reverse the direction of tuning, come to mind when we are faced with this problem in connection with converters. You may think of more as you get involved with this method of tuning.

Circuit Description

Overall, the new MK III seems to work better than the original did. Although no diode selection was made, tracking seems to be as good as that using a conventional 3-gang variable capacitor. Different sets of diodes picked at random were used in the breadboard and the finished PC board with equal results.

A few minor changes were made in the updated version of the receiver. A Motorola MFE521 transistor is used in the rf stage in place of the 40841. A 723 IC regulator is also new, replacing the MFC6030 formerly used. A couple of resistors have also been added to enhance matching to the Murata filters used for i-f selectivity. Everything else is the same as the original.

With the values shown, the

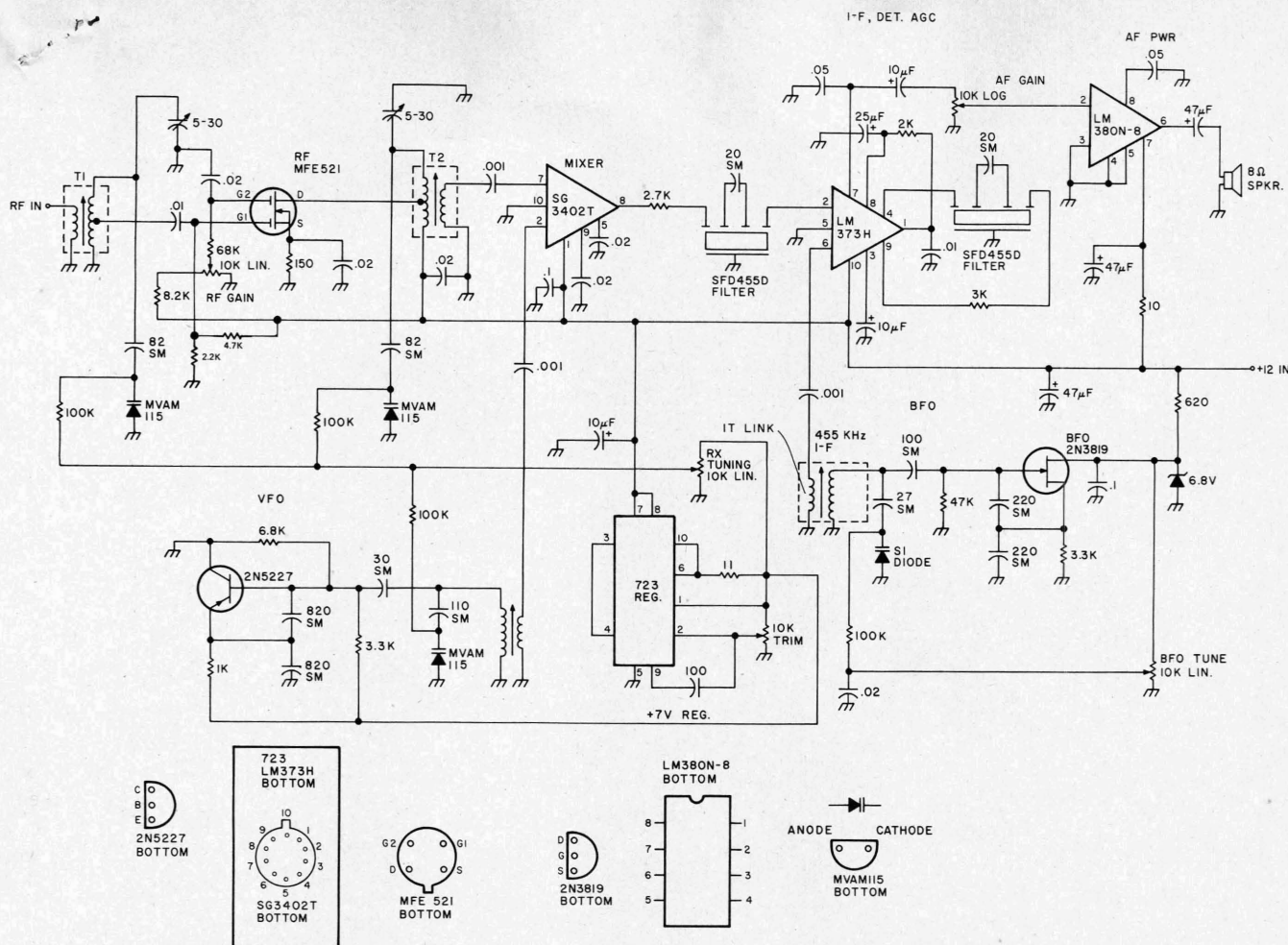


Fig. 1. Complete receiver schematic. All resistors are $\frac{1}{4}$ Watt, 5%. All polarized capacitors are dipped tantalum. All SM capacitors are silver mica. Remaining capacitors are low-voltage discs. Whole numbers are pF unless marked otherwise. Decimal values are in μ F.

tuning range of 3.5 to 4.0 MHz is covered in 180 degrees of rotation of the tuning pot shaft. This makes it easy to use conventional dials.

Because the tuning voltage must be very stable, a voltage regulator is used in place of a zener diode. A 3-volt input-to-output differential is required by the regulator, so, with 7 volts out, the input must be kept above 10 volts. With a 723, the minimum output with the configuration shown is about 7 volts, which, fortunately, is adequate for our needs. The regulator also powers the vfo.

Construction

There are 3 coils or transformers to wind and one i-f transformer to be modified before construction begins. If you've read other Minicom

articles, you know I use standard $\frac{3}{8}$ "-square 455 kHz transistor i-f transformers for a lot of my construction work. In this case, T1 and T2 are wound on stripped-down i-f transformers using salvaged wire. Refer to Fig. 3 for winding instructions. The vfo tank coil is pie-wound on a slug-tuned form. The bfo tank consists of a standard transistor i-f transformer whose secondary has been modified to one turn. This can be accomplished by gently breaking the secondary leads right close to the core with tweezers and unsoldering the remaining wire at each pin. Wind a new one-turn link over the existing windings, and solder the ends to the same pins used by the original winding.

Since it was all being redone, I thought I'd make

this version smaller by mounting the resistors and diodes hairpin fashion. The resulting layout is 2.1" wide by 3.9" long, almost a half inch narrower than the old receiver.

All resistors are $\frac{1}{4}$ -Watt, 5% units with one lead bent around a full 180 degrees so that they can be mounted vertically in closely spaced holes. Diodes are mounted similarly. All polarized capacitors are dipped tantalum. All other capacitors are low-voltage discs, except where silver micas are indicated.

The two 5-30 pF trimmers are very small 5 mm types. These, as well as some of the other parts, may give you trouble when searching for sources. If you send me an SASE and a list of your needs, I'll let you know what

spares I have and what they cost.

After all parts have been mounted, you'll have quite a few empty holes left over. These are meant for connecting leads to external controls and for dc power. The PC layout shows where everything goes. Some holes will not necessarily be used, such as all grounds located around the copper border. These are for convenience and not specifically assigned. You will also find spare pads for +12 volts and +7 volts regulated.

Testing and Operation

With a 12-volt supply, the receiver should draw between 50 and 60 mA with no signal. A meter in series with the power lead or a metered supply can be used for the initial smoke test in order to

Now Try 1296 MHz

—simple discone antenna

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Norfolk VA 23508

Norman V. Cohen WB4LJM
7719 Sheryl Drive
Norfolk VA 23505

Ask most amateurs why they do not try the microwave bands, and you will receive four standard answers:

1. Nobody to talk to.
2. No commercial

equipment available.
3. Communication is limited to line of sight.
4. Construction requires lathe and mill precision work.
The first reply is self-

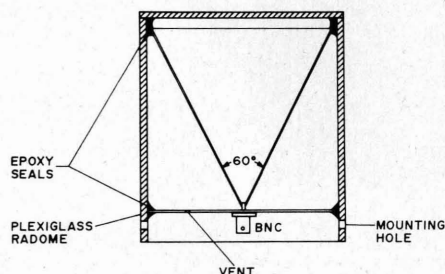


Fig. 1. Inverted discone antenna.

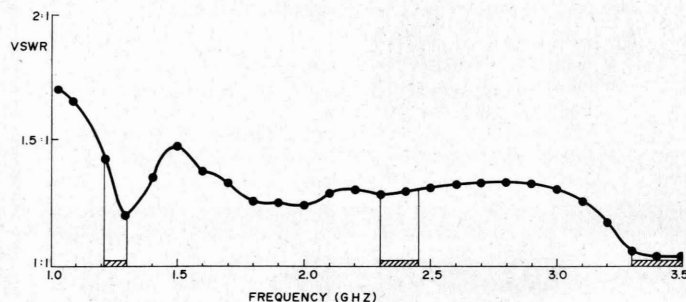


Fig. 2. Vswr of 100 mm inverted discone antenna.

serving — if you get on the band, then there is another operator to talk to. The second reply is sadly true, with the exception of the Microwave Associates "Gunnplexer." The third reply reminds us of the response to two meter FM before repeaters came on the scene. The

fourth is a myth — many of the pioneers in microwaves started with a soldering iron and tin snips. Good examples of what can be done with simple tools are the many fine construction articles by Bill Hoisington K1CLL.

This article shows how to construct simple and efficient broadband antennas for 1296 MHz and up. The construction is not difficult and can be done with simple hand tools. The antennas can be classified as inverted discones. You can find the theory of operation elsewhere; this will be concerned with the practical construction.

Fig. 1 shows a cross section of the inverted discone. The conic portion has a 60-degree apex angle. The cone's base and sides are all made three-eighths wave long at the lowest frequency of interest. Most texts specify this length to be one-quarter wave, but we have found that three-eighths wave gives a significantly lower vswr. The disc diameter is also three-eighths wave. The cone is made by cutting a semicircle

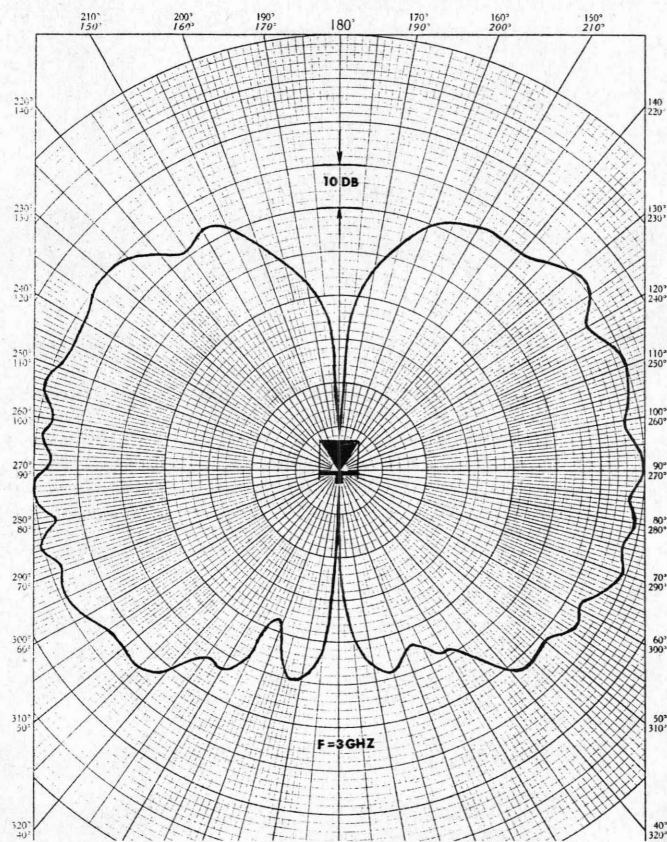
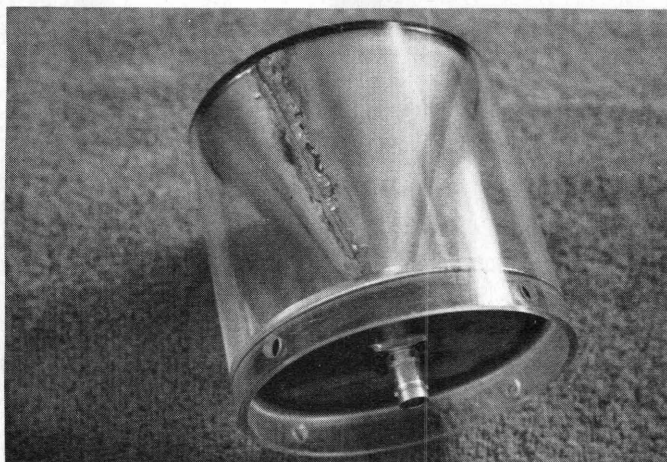


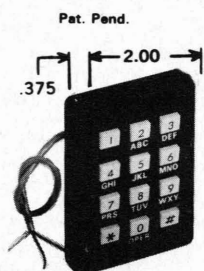
Fig. 3. Polar plot of 100 mm inverted discone.

from copper or brass sheeting with a radius equal to the desired length and rolling it into a cone. The edges are sweat soldered. The BNC connector is sweat soldered to the disc, and then the cone peak is soldered to the connector's center pin. The entire assembly is slid into a Plexiglas™ radome which serves to support the cone and weatherproof the antenna (see the photograph). If the radome is allowed to extend below the disc, it can

serve as a convenient mounting ring. The antenna in the photograph has a base diameter of 4 inches (100 mm). The vswr from 1 GHz to 3.5 GHz is shown in Fig. 2. A typical radiation pattern is shown in Fig. 3. The antenna is vertically polarized. The horizontal radiation pattern is a circle. The useful bandwidth for an inverted discone is 5 to 1; a 1 GHz design is useful to 5 GHz. A smaller unit would be usable to still higher frequencies. ■



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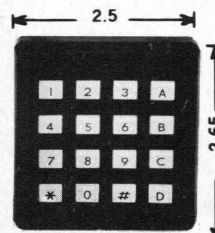
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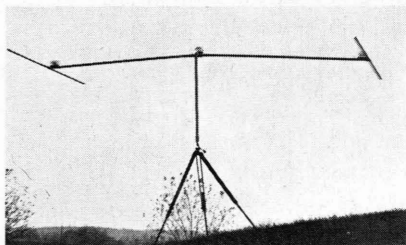


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There are times when one might run into the problem of a specific spurious frequency being radiated by a transmitter and causing interference problems, or reception problems caused by a specific undesired signal. The usual approaches to these problems are various types of LC traps, filters, etc. However, there is another approach which is often very useful as well as economical. No involved construction of tuned circuits is necessary, and one can achieve 30 dB or more suppression of an undesired frequency in either the transmitting or receiving mode.

The basic idea of coaxial stub filters is very simple. As shown in Fig. 1, a section of transmission line which is $\frac{1}{2}\lambda$ -long and which is shorted at one end will reflect an electrical short to the other mechanically-open end of the trans-

mission line. This will only occur at the frequency for which the transmission line is $\frac{1}{2}\lambda$ -long. So, if we have a transmission line going to a transmitter or receiver as in Fig. 2(a), and wish to suppress some frequency, we can insert a $\frac{1}{2}\lambda$ -long rejection stub for the frequency to be attenuated. At that frequency, the coaxial transmission line to the receiver or transmitter will electrically act as if it were shorted. Although there are some cases where only a rejection stub can be used, usually it has to be combined with a correction stub as shown in Fig. 2(b). This is because the rejection stub at the desired operating frequency will present some reactance across the main transmission line. The correction stub cancels this undesired reactance so that at the operating frequency the main transmission line acts electrically as though there

were no stubs at all on it.

That is all there is to the theory of a coaxial stub filter. The rest of this article is mainly concerned with the practical aspects of determining the lengths of the stubs and special applications. The stub filter idea can be used over a very wide frequency range. The limitations to its application are mainly mechanical, in that at very low frequencies the rejection stub may become too long, and at UHF frequencies it may not be possible to dimension the stubs accurately enough. As was mentioned before, the rejection stub has to be $\frac{1}{2}\lambda$ -long at the frequency attenuated. The $\frac{1}{2}\lambda$ length must be an electrical length, so when you're

dealing with coaxial cable it will differ from the physical length because of the velocity factor of the cable. Fig. 3 presents the actual physical length that a coaxial stub should have in order to be electrically $\frac{1}{2}\lambda$ -long at a given frequency. The graph is based on a velocity factor of 0.66, which is the usual one for all common coaxial cables (RG-8, -58, -59, etc.). The graph only covers the frequency range from 10 to 150 MHz, but one can calculate the length of a coaxial stub needed at any other frequency from the formula:

$$\text{Length (inches)} = \frac{3897}{F \text{ (MHz)}}$$

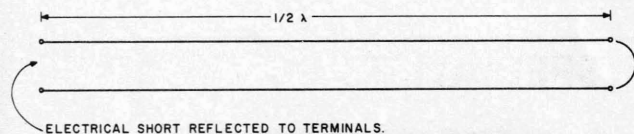


Fig. 1. Stub filters are based on characteristics of a $\frac{1}{2}\lambda$ transmission line.

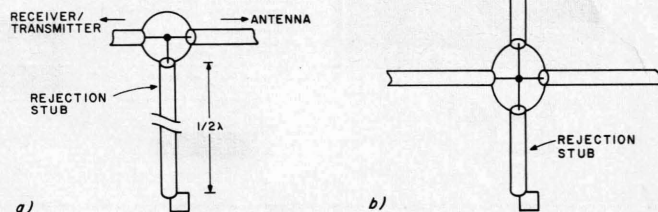


Fig. 2. Placement of rejection stub (a) and rejection stub with correction stub (b). The ends of both stubs are shorted. Although for clarity the shields are shown connected together by jumpers, they should, in reality, be directly soldered or otherwise connected together.

Remember that the frequency involved in this case is the undesired one which one wishes to attenuate, and not the operating frequency.

In practice, one should cut the piece of coaxial cable several inches longer than called for to allow for pruning. Install the rejection stub reasonably close to the antenna terminals of the transmitter or receiver being used. Various types of coaxial connectors are available to interconnect the lines, but the interconnections can simply be soldered on without connectors for indoor usage. Using a pin or pick, short-circuit the outer shield of the rejection stub to the inner conductor at various points starting from the outer end of the stub until maximum attenuation of the undesired signal is achieved. When the best point is found, cut off the excess cable, fold the outer shield around the inner conductor, and solder them together.

The length of the correction stub will vary from zero to $\frac{1}{2}\lambda$. Its length can be determined approximately from the graph of Fig. 4. In this case, by knowing the ratio of the undesired frequency to the operating frequency, one can determine the electrical length of the correction stub at the operating frequency. The physical length of the coaxial cable used can be determined from the formula previously given for a $\frac{1}{2}\lambda$ -length stub. As in the case of the rejection stub, the correction stub should initially be made longer than calcu-

lated. The correction stub is then experimentally shorted until the least attenuation is produced at the operating frequency. If a transmitter is being used, the correct stub length is easily determined by pruning the stub until an swr meter indicates a flat line.

As an example of the foregoing, say we were operating on 21 MHz but had problems with an interfering signal on 26 MHz (in the transmitting case, our operation on 21 MHz was producing some spurious radiation on 26 MHz). The length of the 26-MHz rejection stub would be approximately 150 inches. 26 MHz is about 1.25 F on the graph of Fig. 4, so the approximate length of the correction stub is $1/16\lambda$ at 21 MHz, or about 23 inches.

All of the foregoing may seem like a bit of work, but really all it takes is patience and a slight amount of math. The results that can be achieved by the simple application of stub filters are shown in Fig. 5. Those are not bad results for just the cost of lengths of coaxial cable.

There are several special cases that one should be aware of. If the frequency which is to be attenuated is $1/8$, $1/4$, or $1/2$ of the operating frequency, the idea of coaxial stub filters will not work. This is because the rejection stub will turn out to be a wavelength or a multiple thereof at the operating frequency. Therefore, it will present an electrical short across the transmission line also at the operating frequency. Note that the

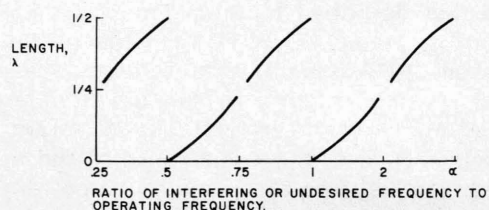


Fig. 4. Graph for determining length of correction stub.

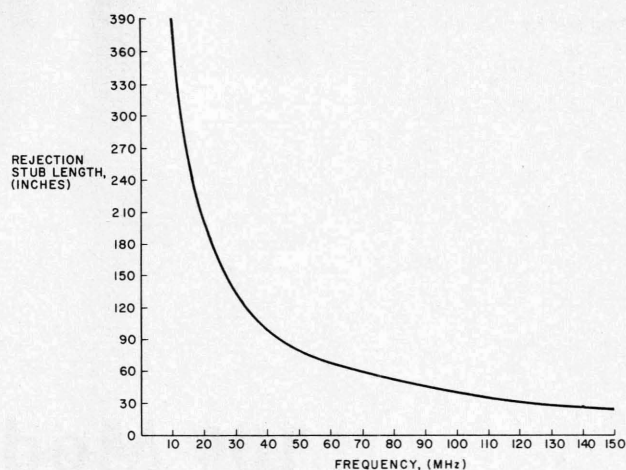


Fig. 3. Graph to determine approximate length of rejection stub.

graph of Fig. 4 is broken at several points, e.g., $.75 F$ and $2 F$. In these cases, the indicated length of the rejection stub would be $\frac{1}{4}\lambda$ at the operating frequency. But, a shorted $\frac{1}{4}\lambda$ stub represents an open circuit, so there is no need for the correction stub at all in these cases. The $2 F$ case is interesting, of course, since it would represent the second harmonic of a transmitter. By using only a rejection stub, therefore, one could get a nice bit of additional second harmonic attenuation. In fact, it will attenuate all even harmonics.

One can use more than one stub filter or different stub filters on various

bands with a switching arrangement. Many combinations are possible, and in some cases only the correction stub may have to be switched on different bands. As a bonus feature of the stub filter idea, note that the shorted end of the rejection stub can be grounded if desired. Thus, one can provide a direct electrical connection to ground for lightning protection, static drain, etc., for an antenna system that does not have to be switched in or out when operating. In a multiple antenna system, one could have a grounded stub for each antenna so that each antenna is continuously protected. ■

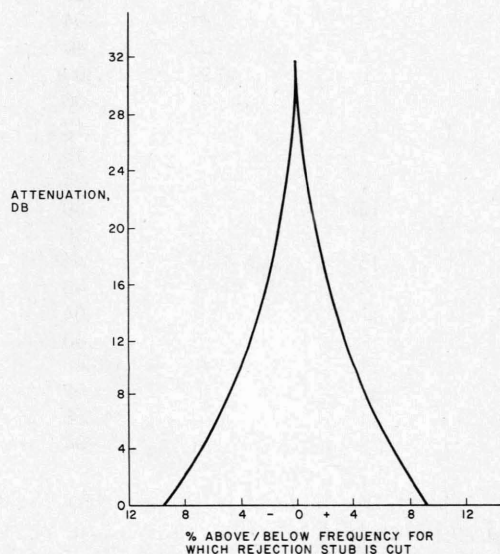


Fig. 5. Attenuation characteristics of typical rejection stub.

You Can Watch Those Secret TV Channels

— a complete MDS receiving system

Good-bye, commercials!

*Jim Barber KØJB
Rt. 1, 22518-97th Ave. North
Rogers MN 55374*

*Jevon Lieberg KØFQA
Rt. 1, 12285 Genereux Place
Rogers MN 55374*

Did you know that there are two secret TV channels? Nobody advertises them, and you can't even buy a TV set that has these channels.

How long have you been complaining about all the commercials while watching your favorite program

or a late night movie? Well, here is the answer to your prayers—these channels *don't even have commercials!*

The programming on these channels consists of movies (P-, PG-, and R-rated), nightclub acts, and sporting events. They

are allocated to Multipoint Distribution Service (MDS). The existence of these channels was written up in 73 last November.¹

If you have heard of MDS via other amateurs, friends, or magazine articles, your curiosity has probably urged you to be on the lookout for a receive system you could build yourself. If this is true, read on!

The MDS Receive System

In this article we will give complete construction details on how to build a cheap and simple MDS receive system. This system will include the antenna, mixer, local oscillator, i-f amplifier, power supply, and complete mechanical layout.

The frequencies of the two microwave MDS video channels are 2154.75 MHz for channel 1 and 2160.75 MHz for channel 2. The audio is 4.5 MHz below the video. For more detailed information about microwave TV, read *A Vidiot's Guide to Microwave TV* by Paul Shuch.²

Locating the MDS Transmitter

If you have seen or

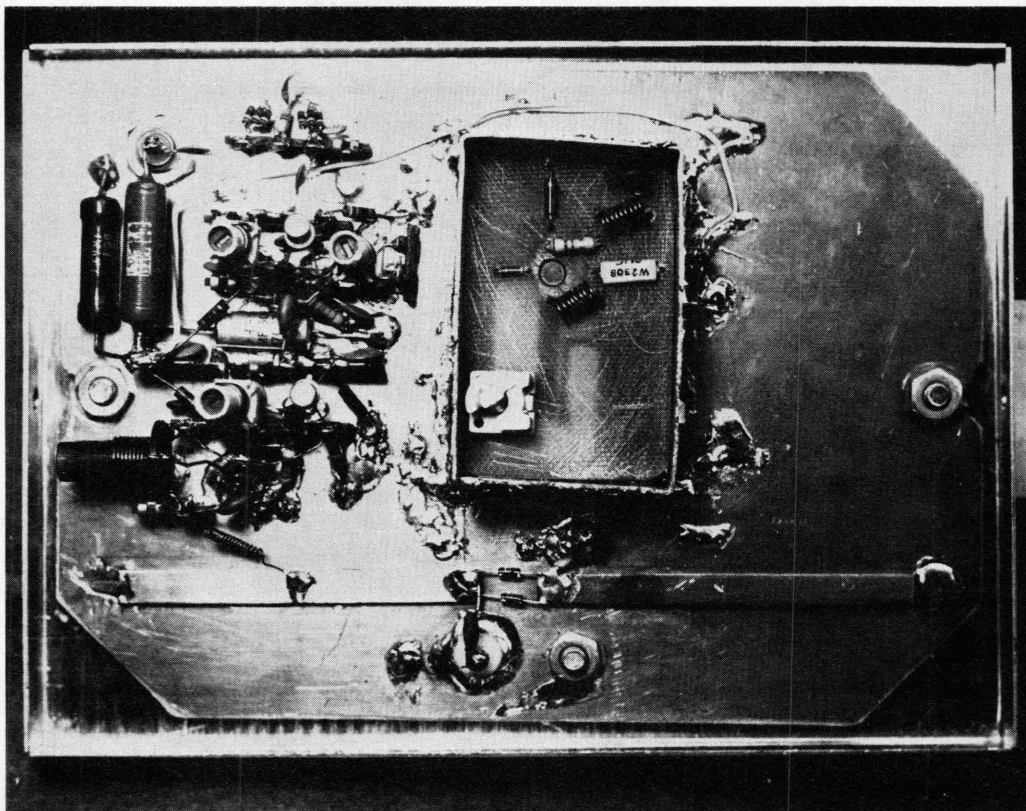


Photo A. This is a close-up of the downconverter showing i-f amplifier, mixer, and the local oscillator in its brass box with the cover removed. The piece of angle aluminum used to mount the box to the mast can also be seen.

heard of small dish antennas located on apartment buildings or hotels, this indicates that there is an active MDS channel nearby.

There are several ways to locate the source of the MDS signal in your area. If your city has an FCC office, they will have a list of all allocated frequencies and the locations of the transmitters. This list is on microfiche and is public information. Stations are listed by their frequency, so look up both MDS channels. The location will be given by latitude and longitude; with a good map, you will be able to pinpoint it.

Another way is to note the direction of several receiving antennas on buildings and use triangulation to find the transmitter. If you can find out the name of the company offering MDS service locally, just call and ask them. One thing is certain—it will be located on one of the highest structures around. After all, their objective is to be line-of-sight to as large an area as possible.

Checking the Receive Path

After finding the source of the MDS signal, you must confirm your location as being suitable for good reception. As you know, microwave signals generally travel only in a straight line, and obstructions such as trees and buildings severely attenuate the signal strength. As a rule of thumb, if you can see the building or tower where the MDS transmitting antenna is located, no matter how distant, then you have a very good location.

If you are line-of-sight, you will not gain much by increasing the height of the antenna. In other words, don't put your antenna at the top of your fifty-foot tower if you can see the location of the signal

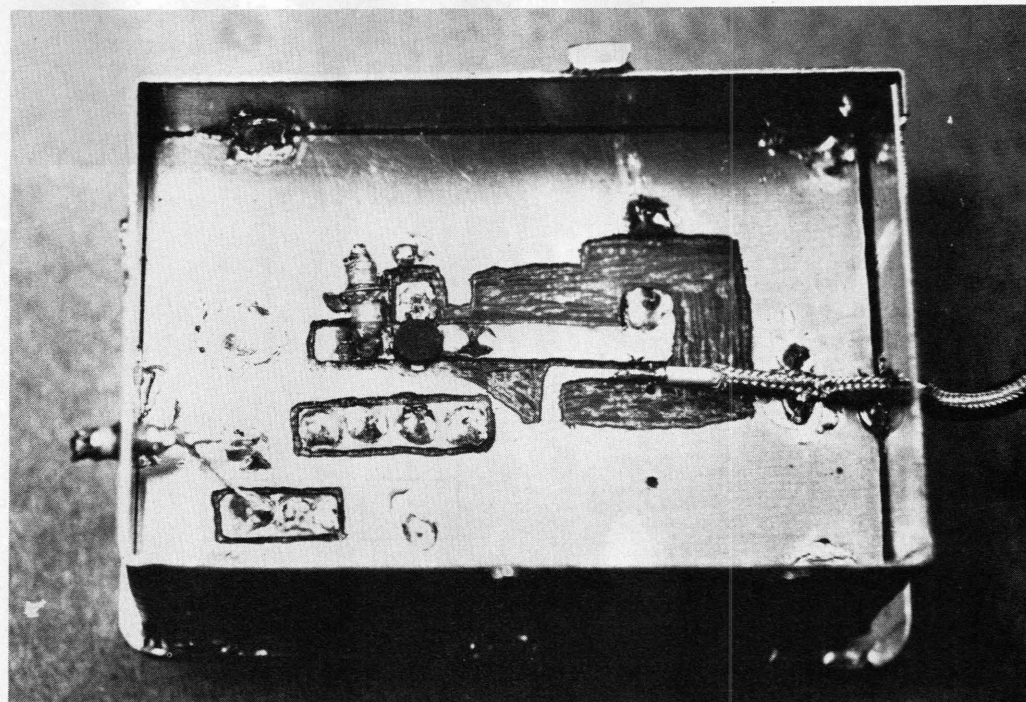


Photo B. This picture shows a close-up of the foil side of homemade printed circuit board used for the local oscillator. Note the modified feedthrough capacitor used for a base bypass capacitor. A 270-pF disc with very short leads also can be used.

source from your living-room window. Some obstructions can be tolerated if not too severe, and the signal loss can be made up through the use of preamplifiers and/or a larger antenna.

Circuit Description: Mixer

The mixer is the secret to a good microwave downconverter. It normally contains most of the critical parts and generally is the most difficult to construct. I think we can speak from experience on this subject, having built the interdigital converter by WA2CQH,³ the high-performance balanced mixer for 2304 MHz by WA2ZZF,⁴ and the solid-state 2304-MHz converter by K2JNG, WA2LTM, and WA2VTR.⁵ Then, along came the October, 1978, *Ham Radio* magazine, with Jim Dietrich and his twin-diode mixer article.⁶

This mixer is so simple that you find it hard to feel that you have earned the quality of the signal that comes out of it. The mixer consists of two half-wave lines mounted above a ground plane. The first

half wave is grounded on one end, and the other end has a pair of back-to-back diodes connecting it to the second half-wave line. This second line is open on its far end, and the i-f output is taken from the center of it.

The rf input is at the diode end of the second half-wave line, and the local oscillator input is at the diode end of the first half-wave line. (See Fig. 1 for detail.) The input circuit at the rf port is a high-

pass filter to keep the i-f energy from getting out of the rf port. This filter also keeps the antenna input at dc ground. The output circuit at the i-f port is a low-pass filter to keep the LO out of the i-f. For a more detailed description of the mixer, read Jim Dietrich's article.⁶

The diodes used in the original article were MA4882. We were unable to get these locally, so we used Hewlett-Packard 5082-2835. This is a very

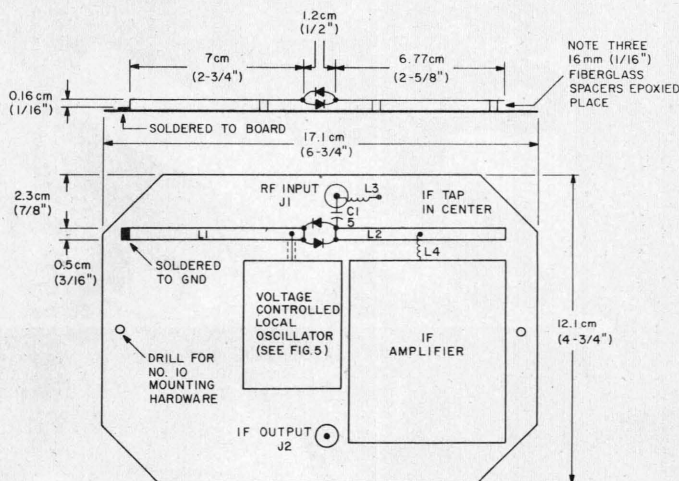


Fig. 1. General layout of the downconverter and construction details of the twin-diode mixer. J1 is a type N or BNC. J2 is an F61 TV-type connector.

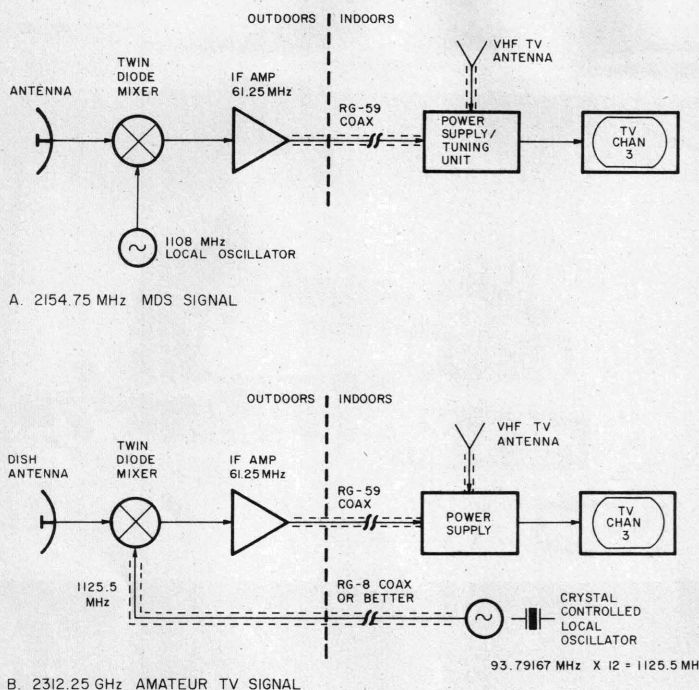


Fig. 2. (a) Block diagram of MDS receive system as described in this article. (b) Block diagram using alternate crystal-controlled local oscillator. Frequencies shown are correct to convert down to standard VHF TV channel 3 (video).

fine inexpensive diode for microwave mixers, but H-P does not give a noise figure rating for it. Any good Schottky diode can be in this circuit, but if you want to know its noise figure rating, you will have to pay more. The 2835 is about

\$1.10, and the prices go up from there. A good compromise, possibly, is H-P's 5082-2817. It has a 6-dB noise figure at 2 GHz and costs only \$1.85.

Block Diagram

The block diagram

shown in Fig. 2(a) is representative of the MDS receive system as described in this construction article. This system consists of a mixer, voltage-controlled local oscillator, and i-f amplifier located at the antenna, and the com-

bination power supply/tuning unit, located at the TV set. The RG-59/U feedline then performs the dual purpose of carrying the i-f signal down the line to the TV and carrying the power supply/tuning voltage up the line to the converter.

This results in a nice, cheap, and efficient system requiring no expensive feedline and connectors and a minimum of critical circuits. No special test gear is required for tune-up.

The block diagram shown in Fig. 2(b) is one of many configuration alternatives possible and is shown mainly to provoke some thought about other uses for the basic system. An example of another use would be for fast-scan amateur television where the usual DX signal would be much weaker than an MDS signal, and the guesswork of tuning is eliminated through the use of a crystal-controlled local oscillator.

Circuit Description: The Local Oscillator

Building a satisfactory local oscillator is the next hurdle one must overcome in the construction of a microwave converter. You will be thankful to know that the twin-diode mixer helps us here, too, because this mixer requires an LO running at only half the frequency needed for other mixers.

In this system we have done away with all the drudgery of those crystal-controlled oscillator/multiplier chains also and settled on a nice simple free-running oscillator.^{8,9} This oscillator has a tuning range of about 900-1300 MHz and a power output capability of several milliwatts. You might ask, "Is that possible? Won't it drift?" The answer to both questions is "yes." Not only is it possible, but it's also very simple.

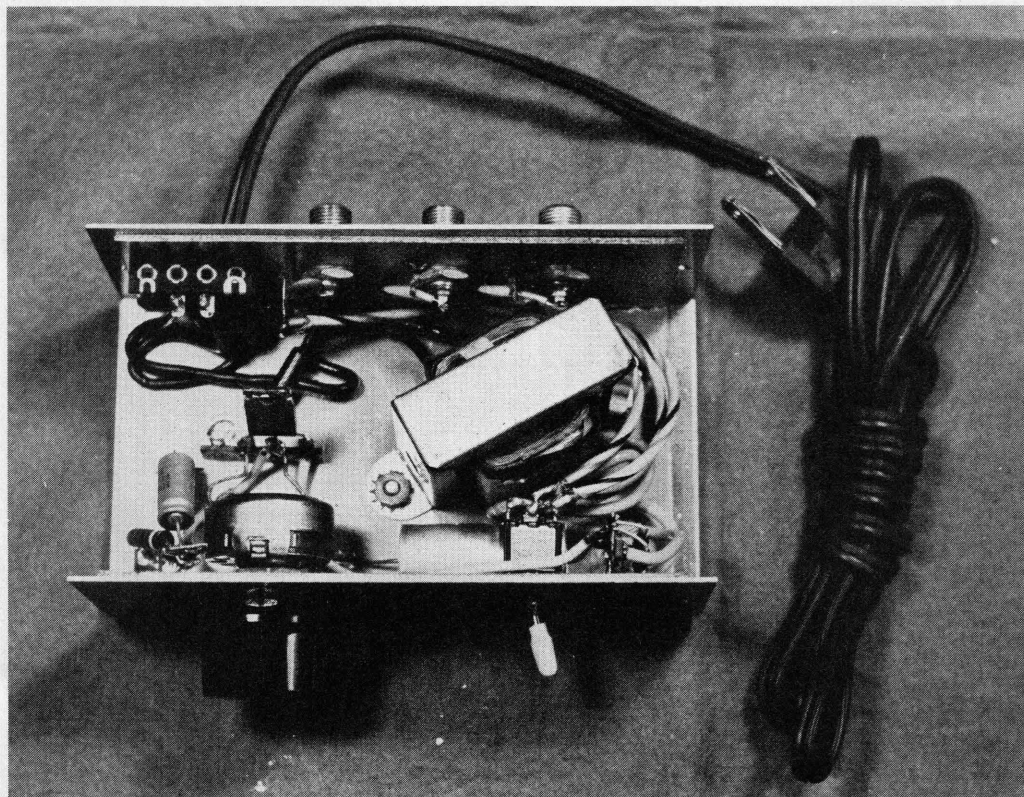


Photo C. This is a top view of the power supply/tuning unit.

The oscillator consists of a printed circuit board with an etched inductor, a Motorola MRF901 transistor, and about six other parts. This is about as simple as it could be. As far as drift goes, we have found that it is at a very slow rate, completely due to changes in outside temperature.

A TV set is a very broad receiver and will tolerate considerable drift. Since this LO is a voltage-controlled oscillator and can be remotely tuned from the control box, frequency adjustments can be made at any time. If this type of stability is not satisfactory for you, you may wish to build a crystal-controlled LO. We might suggest the one used on the interdigital converter³ with a doubler from a Paul Shuch article.¹⁰ Possibly another consideration along this line would be the Paul Wade article in *Ham Radio*, October, 1978.¹¹ Although these alternatives would double the size of the converter, it may be worth it—particularly if you plan to distribute the signal to a number of TV sets. This is because, when tuning the converter for one TV set, the others are affected. This may be necessary once your neighbor finds out what you are watching.

Circuit Description: I-f Amplifier

The i-f amplifier is a two-stage amplifier using dual-gate MOSFETs. The function of the i-f amplifier is to boost the signal from the mixer and also provide enough power to drive the RG-59/U coax line to the control box and TV set. This amplifier has been used to drive up to 600 feet of RG-59/U cable with enough signal at the far end to still have a good picture. Although the center frequency is 63 MHz (channel 3), the amplifier is broad enough to cover channels 2 through 4. The output of



Photo D. This photo shows a complete MDS receive system. The tripod, mast, and U-clamps are standard TV items. All connectors and the box containing the downconverter should be weatherproofed.

the amplifier goes into a 75-Ohm, 3-dB pad. This pad gives the amplifier a constant impedance to look into, and, along with the ferrite beads on gate two of both of the MOSFETs, helps make this amplifier very stable.

Circuit Description: Power Supply/Tuning Unit

The power supply used in the MDS receive system performs two separate functions. First, it supplies the operating voltage for the i-f amplifier and local

oscillator, and second, because it is adjustable and the local oscillator is frequency-sensitive to voltage, it provides tuning control for the downconverter.

The 24-volt ac transformer, diode bridge, and filter capacitor provide about 34 volts dc to the 12-volt regulator. The regulator is made adjustable by the addition of the fixed resistor, R1, and the variable resistor, R2. With the values shown in Fig. 4, the voltage is adjustable

from 12 to 17 volts.

Rf choke RFC1 isolates the rf from the supply, and the dc is carried via the feedline to the downconverter. DPDT switch S1 performs the dual function of switching the TV between the VHF antenna and the MDS downconverter and properly terminating the unused lead with a 75Ω resistor.

It may seem that the current capacity of the power supply far exceeds the requirements of the downconverter, and it does. The

board.

Remove the board from the cover and begin building the twin-diode mixer. Cut the two copper or brass lines as shown in Fig. 1 and attach to the board by soldering the end shown as grounded, and using epoxy cement and three small .16-cm (1/16-inch) fiberglass spacers as shown. Scraps of PC board are used for the spacers. You may have to rough up the copper a bit with emery paper in order to get the epoxy to stick well.

Next, wire the i-f amplifier and complete the wiring of the mixer as shown in the schematic in Fig. 3. Use short, direct wiring, using the copperclad board for all ground connections. The unit shown in Photo A uses point-to-point wiring, using Bakelite™ terminal strips. This seems to work well, and no trouble has been encountered. Don't forget the ferrite beads on gate 2 of the two MOSFETs, and keep L5-8 as far apart from each other as possible.

The local oscillator shown in Photo B is built on a homemade printed circuit board. The board is a 4-cm (1½-inch) by 6-cm (2¼-inch) piece of .16-cm (1/16-inch) G10 single-sided circuit board. A Dremel Moto-tool® with a small carbide tipped burr was used to grind away the unwanted copper plating. Fig. 5(a) shows the circuit board layout and dimensions. The circuit is not really critical, but the board layout should be followed as closely as possible for best results.

Bypass capacitor C2 is a modified 270-pF feedthrough type. Both ends are snipped off as short as possible, then one end is soldered to the base of Q1 and the mounting ring is soldered to ground. Tuning capacitor C1, shown in Photo A, is a Johnson type U 189-502-5 1.3-6.7 pF air-

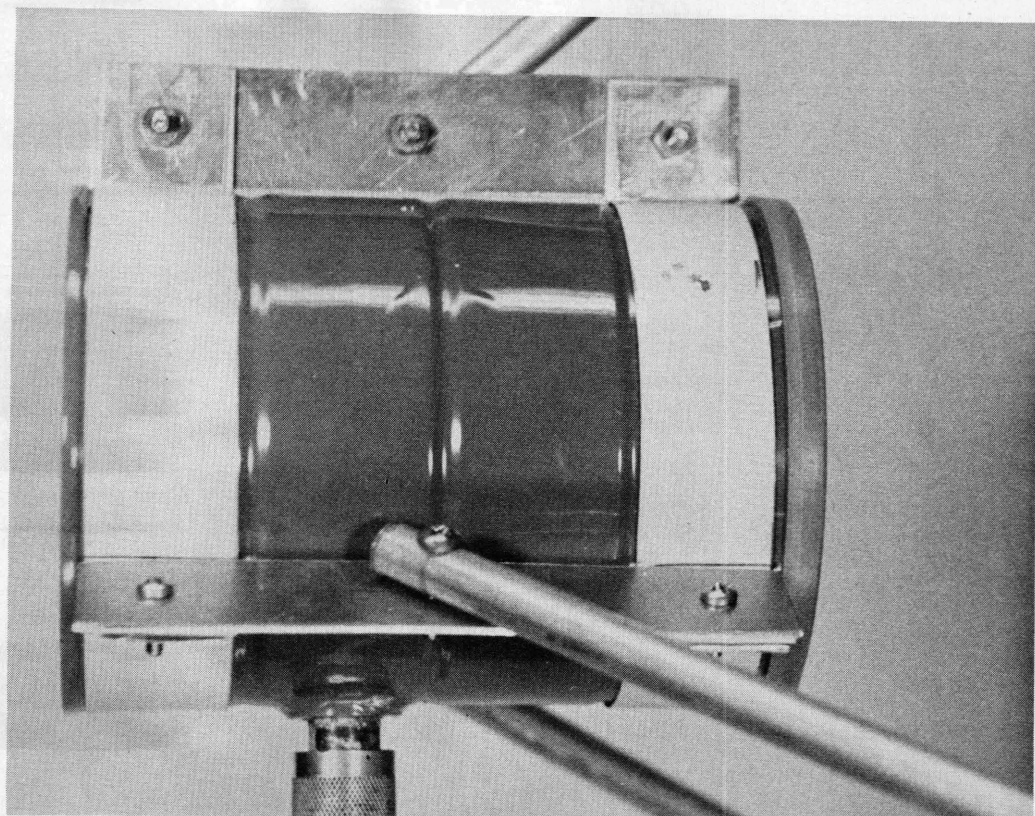


Photo F. Here is a close-up of the completed feedhorn and supporting hardware. Note the plastic radome cover is in place.

variable. Later, it was found that a small tubular type such as a Centralab 829-3 .5-3 pF worked better. Using a tuning capacitor with an NPO temperature coefficient and metal film resistors reduces frequency drift due to temperature changes. Wire the remainder of the local oscillator as shown in Fig. 5(b).

The local oscillator is mounted in a small box fashioned from a strip of brass 3.2 cm (1¼ inch) wide and about 20.2 cm (8 inches) long. The board is suspended in the center of the box by soldering the edges of the circuit board to the center of the inside walls of the box. Holes for the B+ feedthrough capacitor, C3, and rf output coaxial cable are drilled, then the box is placed near the mixer as shown in Fig. 1 and temporarily tack-soldered in place. Make a cover for the box by cutting a 4.5-cm (1¾-inch) by 6.5-cm (2½-inch) piece of printed circuit board or thin brass. Drill a hole in it

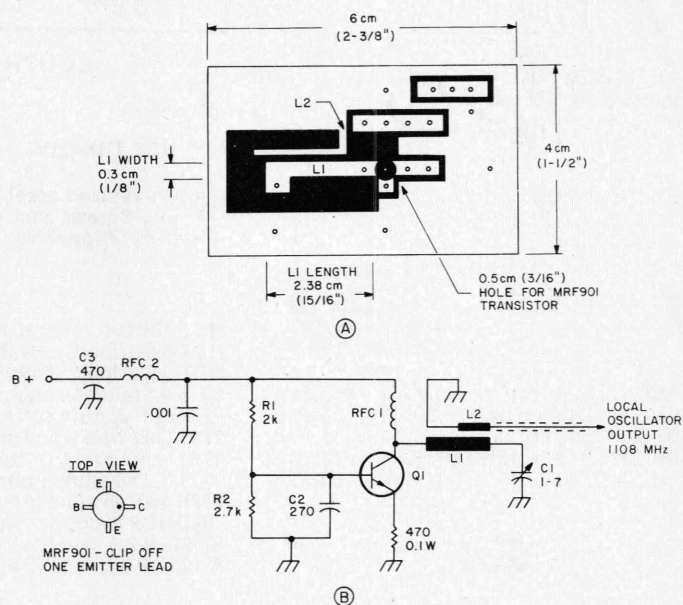


Fig. 5. (a) Bottom-view layout of local-oscillator circuit board. The shaded area indicates the area etched away. Use single-sided G10 epoxy circuit board. (b) Schematic diagram of local oscillator. Q1 is Motorola MRF901. RFC1, 2 is 13.5-cm (5¼-inch) #26 AWG close-wound, using a piece of #14 AWG as a mandrel. C1 is a 1-7-pF Johnson-type U (see text). R1 and R2 are ¼-W metal film-type resistors. C2 is a modified 270-pF feedthrough (see text) or other small low-inductance capacitor. This capacitor is soldered to the foil side of the circuit board to ensure short lead lengths. L1 and L2 are PC board inductors. Keep L2 as narrow and as close to L1 as possible. See Fig. 5(a).

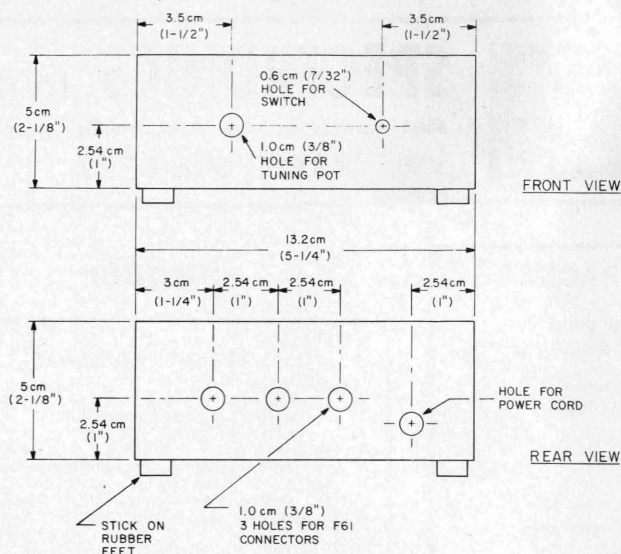


Fig. 6. Front and rear panel layout of the power supply/tuning unit chassis.

to permit access to tuning capacitor C1, and tack-solder the cover in place. Permanent attachment is not done until checkout and tune-up are complete.

Power Supply

The combination power supply/tuning unit is built into a 13.3-cm (5 1/4-inch) by 7.6-cm (3-inch) by 5.4-cm (2-1/8-inch) aluminum box. An LMB Tite-Fit chassis #780 works well. The box is

drilled to accept mounting of the controls and type F61 coaxial connectors. The function switch is a DPDT subminiature, requiring a .6-cm (7/32-inch) hole. Holes for the coaxial connectors, the tuning pot, and the ac line grommet are 1-cm (3/8-inch). See Fig. 6.

The unit shown in Photo C was lined with double-clad printed circuit board. This was done in order to

accommodate soldering to ground conveniently. This is a personal construction preference and certainly is not required. At this time, before the hardware is attached, the two-piece box may be covered with contact paper or painted as the builder desires. Small rubber stick-on-type feet are then attached.

Point-to-point wiring using Bakelite terminal strips was used in the unit shown. Follow the schematic in Fig. 4 and no trouble should be encountered.

Antenna Construction

The antenna design is shown in Figs. 7 through 10 and Photos D, E, and F. For the reflector bracket, you will need two 79-cm (30-inch) pieces of 1.27-cm (1/2-inch) aluminum tubing. Bend both ends of the pieces of tubing 90° to make the two U-shaped brackets as shown in Fig. 7. Use an electrician's 1/2-inch EMT conduit bender for this job, bending very slowly, and you will end up with a nice kinkless bend. Now, flatten about 1.9 cm (3/4

inches) of each end of the two U-brackets in a vise and drill a .5-cm (3/16-inch) hole in the center of each of the flattened areas. Bend these flattened ends back at an angle that matches the contour of the outside edge of the back of the reflector.

Next, cut a 14-cm (5 1/2-inch) by 16.5-cm (6 1/2-inch), by .3-cm (1/8-inch) thick piece of aluminum plate. Drill a .5-cm (3/16-inch) hole in each corner and the four .64-cm (1/4-inch) holes for the TV U-clamps. Center this plate on the two U-brackets and mark and drill four .5-cm (3/16-inch) holes in the U-brackets, matching the corner holes in the plate. Attach the plate to the brackets using four #10 bolts, nuts, and lock washers. See Fig. 7.

Now, hold the bracket against the back of the reflector, mark, and drill four .5-cm (3/16-inch) holes in the reflector, matching the holes in the flattened ends of the U-bracket. Attach the bracket to the reflector using four #10 bolts, flat washers, lock washers, and nuts.

The feedhorn is made from a one-pound coffee can measuring 10.2 cm (4 inches) in diameter by 13.7 cm (5-3/8 inches) in length. Drill a 1.3-cm (1/2-inch) hole 5.1 cm (2 inches) from the bottom of the can. Prepare a flange-mount type N or BNC connector (remembering that this is a microwave antenna, and SO-239/PL-259 connectors will not work) by soldering a 2.9-cm (1-1/8-inch) by .5-cm (3/16-inch) piece of copper or brass tubing to the center pin. Solder this connector into the hole previously drilled in the can. Soldering the connector to the can is recommended for both electrical and mechanical integrity. (See Fig. 9.)

Before mounting the feedhorn to the reflector, it

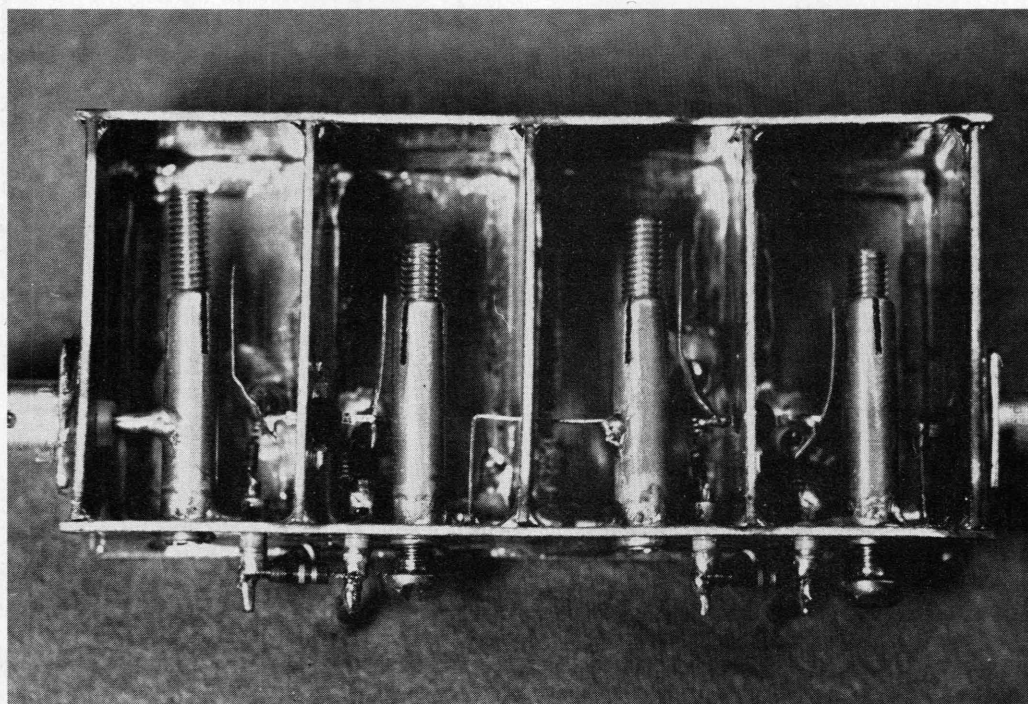


Photo G. Here is a top view of a dual version of the WA9HUV 2304-MHz preamp. This preamp covers the MDS band with no modifications. Note the L-shaped coupling link passing through the center partition.

should be given a thorough paint job inside and out. Don't paint the rf connector, the dipole, or the plastic cover. A good quality, anti-corrosion exterior metal paint should be used.

The feedhorn is held over the focal point with a three-legged support. The legs are made from three pieces of 1-cm (3/8-inch) diameter aluminum tubing 38 cm (15 inches) long. One end of each leg is flattened 1.3 cm (1/2 inch) back, and a .3-cm (1/8-inch) hole is drilled in the center of each flattened area. Drill a .3-cm (1/8-inch) hole in each of the legs 1.27 cm (1/2 inch) from the opposite end and 90° around from the first hole. Bend the flattened ends back to about a 45° angle. See Fig. 8.

In order to attach the supporting legs to the feedhorn, two straps are made of thin aluminum flashing material. Two strips, 2.5 cm (1 inch) wide by 46 cm (18 inches) long are bent and drilled, as shown in Fig. 8. These straps are wrapped around the can and tied together, using three 12.7-cm (5-inch) by 2.5-cm (1-inch) pieces of .16-cm (1/16-inch) aluminum sheet. Six 2.5-cm (1-inch) square pieces of the same material are used as washers, and #8 hardware is used to hold it all together. The supporting legs are then bolted between the reflector and the strap ties, using #8 hardware. See Fig. 10.

All of the hardware used in this antenna should be of the plated type to assure long life; you wouldn't want the thing to fall apart while you're watching something good on the tube.

The open end of the can faces the reflector, and, for proper focus, should end up 28 cm (11 inches) from the surface of the reflector.

All of the dimensions used here are correct for

the 63.5-cm (25 inch) aluminum saucer sled. Some adjustments may have to be made if a different reflector is used.

Care should be taken in choosing a mount for the antenna. The reflector presents a surface area of over .3 square meters (3 1/2 square feet) to the wind. A 61-cm (2-foot) roof tripod fitted with a 3.2-cm (1 1/4-inch) mast should be able to withstand all but the strongest winds. Mount the antenna as low as possible above the tripod, and make sure all bolts and clamps are secure. If possible, mount the antenna in a wind-sheltered location that still has a line-of-sight view of the MDS transmitting antenna.

Final Assembly and Tune-Up

Attach the antenna and downconverter to the mast and set in a tripod as shown in Photo D. Connect a short piece of good coax between the feedhorn and the downconverter. We

have found that Berk-tek brand ultra-flex RG-8X coax fitted with BNC connectors works well for this purpose. The remainder of the cabling uses RG-59/U coaxial cable and TV-type F connectors.

The system can be tuned up with little or no test gear. If you are lucky enough to have a signal generator that covers 2.15 GHz, by all means use it. But, if you are like the rest of us, you will have to get by without it.

The tune-up steps will be to first check out the power supply, tune the oscillator, and peak up the i-f amplifier. This tune-up requires a location that is line-of-sight to the MDS transmitter, easy access to the downconverter and antenna assembly, and a TV set.

Don't attempt tuning up the downconverter on your roof or tower. Play it safe!!

Ensure that the power supply is operating properly by measuring the output voltage. It should be

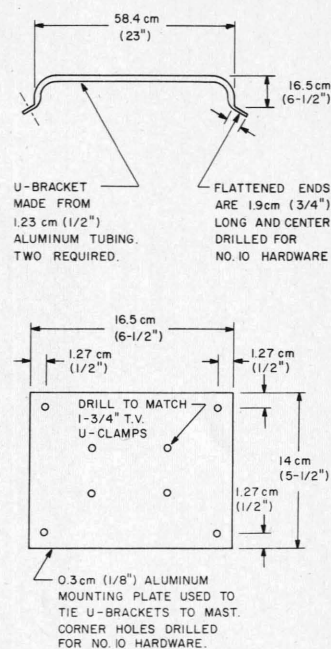


Fig. 7. Construction details of the U-brackets and mounting plate.

adjustable from 12 to 17 volts by varying the tuning pot, R2. This check should be done before applying power to the downconverter.

I-f Amplifier Tuning

Tuning the i-f amplifier is

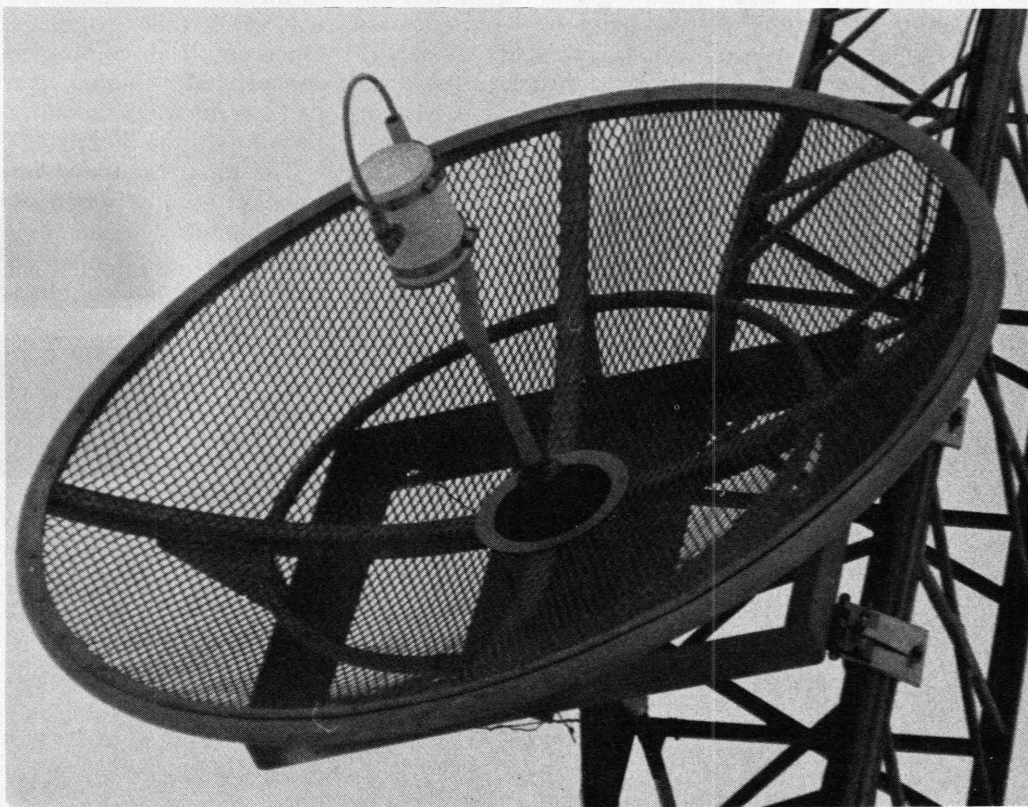


Photo H. This is a 122-cm (4-foot) commercial dish using a WA9HUV coffee can feedhorn. This antenna is used for MDS reception in a marginal signal location.



Photo 1. This is a 74-cm (29-inch) commercial reflector using a WA9HUV coffee can feedhorn. Located 4.5 meters (15 feet) high, this antenna pulls in snow-free color MDS TV signals over a distance of 32 kilometers (20 miles).

quite simple. First, you must choose which TV channel you will be using in your system. This should be channel 2, 3, or 4. It should not be in use in your area. Inject a signal into L5 from a signal generator or a grid-dip meter. The signal frequency should be 57

MHz for channel 2, 63 MHz for channel 3, or 69 MHz for channel 4. On the output of the i-f amplifier, use an rf probe or a series diode and a .001-uF capacitor to ground. Peak all coils (L5, L6, L7, L8) for maximum voltage across the .001-uF capacitor.

If you do not have a signal source at the i-f frequency, then set the slugs in all of the i-f coils halfway into their windings. Once you have completed the oscillator tuning, you can use the MDS signal for peaking up the i-f coils.

The local oscillator frequency required when using channel 2 as an i-f is 1105 MHz, for channel 3 it is 1108 MHz, and for channel 4 it's 1111 MHz. Set the power supply tuning control to midrange. Set switch S1 to the MDS position. Using an insulated

tuning wand, adjust C1 to the proper frequency. If you have a frequency counter or spectrum analyzer that covers the LO frequency, use them. We have found, however, that using an off-the-air signal and adjusting for a good TV picture is the fastest and most sure way to accomplish this.

While tuning C1, you will probably see all sorts of strange patterns on your TV screen along with two or more settings that produce a standard TV picture. The tuning range of the LO is enough so that both the signal that we want and its image can be tuned in. Tune C1 to the higher of the two, which is the setting when the plates of C1 are the least meshed. The tuning control in the power supply can now be used for fine tuning. In the event that C1 shows no effect, you can test the oscillator for activity by using the signal sniffer shown in Fig. 11.

Preamps

There undoubtedly will be some locations that will have marginal signals. If, after having built the system, you find that your signal is snowy, you may wish to improve it with a preamp. There have been several good articles about 2304 preamps in the ham publications in recent years. The one that appears easiest to build is in the July, 1974, *Ham Radio*, by WA9HUV.¹⁵

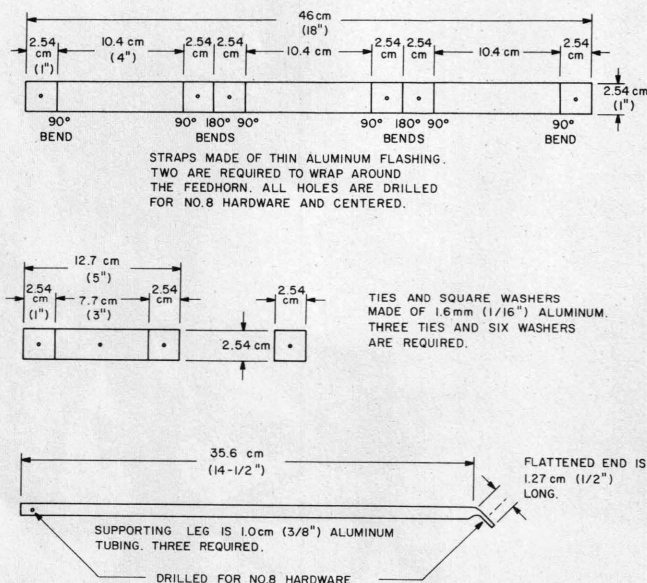


Fig. 8. Construction details of the feedhorn mounting straps, strap ties, and support legs.

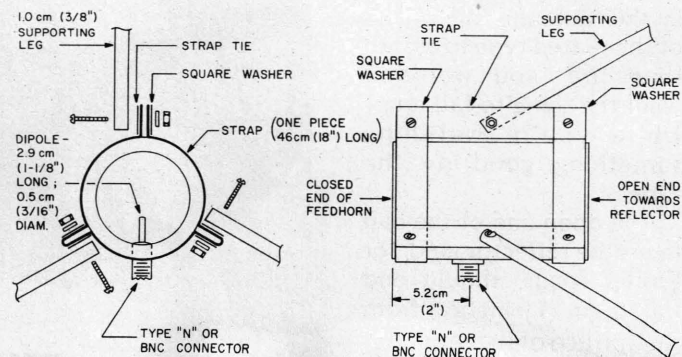


Fig. 9. Feedhorn mounting details.

You will find that most of the transistors the article recommends run from \$17.50 to \$70.00 each. As you can see, this could get expensive, especially if you need two or three preamps in series for more gain. What we have done is change the transistor to a MRF901, which costs \$1.44. This device has less than a 3-dB noise figure at 2 GHz, but only about 6 dB of gain. Now, 6 dB of gain when you have a snowy picture does not help much. For a location that is marginal, we have built two of these preamps in the same box. That gives about 12 dB of gain and also eliminates the need for two connectors. At this frequency, too many connectors in the line can add a lot of loss and is expensive. I am sure it would be possible to build more stages into one box if needed, but we have not yet tried this. We have had up to three of these double preamps together with short cables. This worked well and gave a lot of gain.

We built most of our preamps using double-sided printed circuit board. This is faster and much easier than brass. Be sure to solder together intersecting edges on the inside where possible. At this frequency, a good rf-tight box is desirable. WA9HUV's article gives two possible circuits depending on what kind of transistor is used. When using a MRF901, use the circuit on page 8 of the article, which is labeled

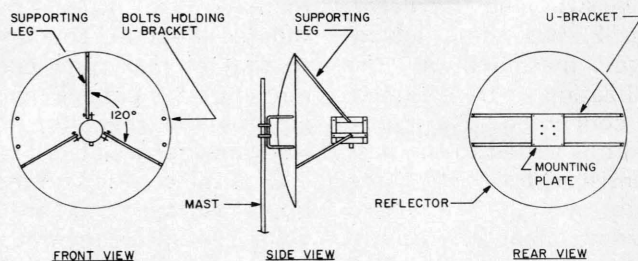


Fig. 10. Final construction of saucer sled antenna. The open end of the feedhorn should end up 28 cm (11 inches) from the surface of the reflector for proper focus.

Fig. 1. We show the method of coupling two stages together in Photo G here.

The preamp should be mounted as close to the feedhorn as possible, and it also will need to be enclosed in a weatherproof box. Power tapped from the downconverter can be used to power as many preamps as are found to be necessary.

Conclusion

In this article we have shown you details for building a cheap and easily-reproducible MDS receive system. Most of the circuits and ideas have been gleaned from previous construction articles and manufacturers' applications notes, so no originality is claimed. Although the system as described performs very well, if you find that you can make any significant improvements to the systems, we certainly would be interested in hearing of them.

Since the original design of the system, printed circuit boards have been laid out and are now being made. These boards, along with some of the harder-to-find components and complete antennas, are available from the authors. Send an SASE for details. ■

Acknowledgements

We would like to thank John Fox W0LER and Ernie Simon W9JCE for the help and encouragement they gave in completing this project.



Photo J. Here are snow saucers before modifications. Shown are the steel 69-cm (27-inch) model made by Flexible Flyer and the 64-cm (25-inch) aluminum Sno Coaster made by Mirro. Remove the handles before using as a reflector.

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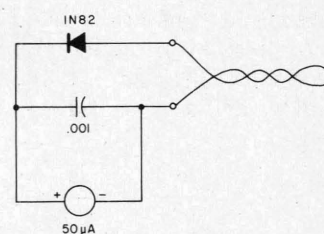


Fig. 11. A piece of test equipment called a signal sniffer. It can be used to tell if the oscillator is working. The meter movement should be a good quality 50-uA movement. The diode is a 1N82 or other high-frequency diode, and the pickup loop is about twelve inches of #22 hookup wire twisted together with a one-inch loop at the end. This sniffer works to 2 GHz or more.

Try the Potted J

— a 2m antenna impervious to the elements

Structural integrity and simple construction.

Some antenna is better than no antenna at all. In many circumstances where adverse weather conditions are encountered, a primary consideration should be to have an antenna that will function dependably even though it might not provide the last dB of gain. This article describes such an antenna designed primarily for 2 meter repeater usage. The antenna has only a modest amount of gain over a simple ground-plane antenna.

But, it has far better structural integrity. The materials used to construct the antenna might vary a bit depending on what is available locally, but the design permits construction by anyone using only simple hand tools.

The antenna form is a commonly used variation of the old-fashioned J antenna shown in Fig. 1(a). The variation, as shown in Fig. 1(b), simply uses a closed metal cylinder for the lower $\frac{1}{4}\lambda$ section in-

stead of the open-style construction of the original J antenna. This type of construction has a number of advantages, such as easy mounting on any type of mast, relatively inconspicuous appearance, and an all-metal, dc grounded structure. Since the antenna's central element and the metal cylinder are electrically shorted at the base of the antenna, that point will be a low impedance point. At the other end of the $\frac{1}{4}\lambda$ cylinder, there will be a high impedance point. The latter is the only point where consideration has to be given to proper insulation between the metal elements of the antenna.

The mechanical dimensions of the antenna are given in Fig. 2(a). Depending on the materials available locally, the diameters of the cylinder and the radiating element can vary a bit from those shown. But, the lengths should be

closely maintained and dimensioned using the formulas shown in Fig. 1(a) for the particular segment of the 2 meter band of interest. The mechanical construction can vary a bit from that shown as long as the dimensions are maintained, the central element is securely grounded to the bottom of the $\frac{1}{4}\lambda$ cylinder, and the coaxial feedline is connected properly.

In the method of construction shown, the end of the central element is flattened out at the bottom and bent and bolted to the bottom side of the $\frac{1}{4}\lambda$ cylinder. Additionally, a small piece of the same material as the central element is flattened out and used as a brace to the other bottom side of the cylinder. The coaxial cable shield is connected via a ground lug to the inside of the cylinder and the inner conductor is either soldered or connected to the central element

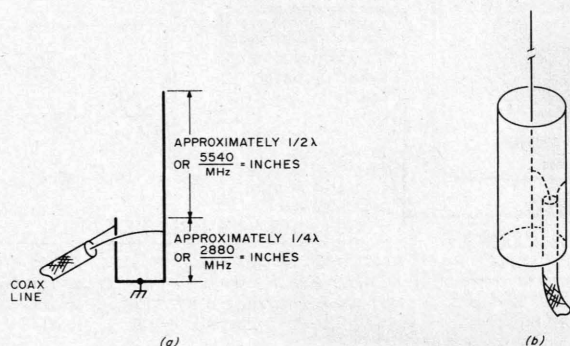


Fig. 1. Basic J antenna (a) and variation where lower $\frac{1}{4}\lambda$ section is in the form of a cylinder (b).

at the point indicated via a ground lug. The connection to the central element can be made before that element is inserted in the cylinder. The connection of the cable shield must be made after insertion. This requires a bit of dexterity working in the narrow cylinder, but with good lighting and patience it can be done. One handy way to help things along is to temporarily glue the nut for the bolt holding the ground lug on a thin piece of wood. Once the nut is started on the bolt, the wood piece can be broken away.

Before the final assembly step, check the electrical performance of the antenna for swr. It should check out with a very low swr if proper dimensions have been maintained. If one wants to optimize the swr, the feedpoint on the central element can be varied slightly up or down. Usually this can be done without having to change the connection point for the ground shield of the coaxial feedline. This does mean going through the procedure of having to take out and reinsert the central element, but it is not at all that tedious after one does it once or twice. During this test, the central element can be held centered in the upper end of the cylinder by a PVC reducer fitting as shown in Fig. 2(b). These reducer fittings can be found wherever PVC piping fittings are available and one can be found which will fit exactly over a pipe having a 1 1/4" outside diameter.

The final and most important step in assembly is to fill the cylinder in completely with an insulating compound. This will give the antenna its final mechanical rigidity and, more importantly, completely exclude moisture leakage or condensation in the cylinder. Many potting or seal-

ing compounds can be used as long as they contain no form of metal filler. The plastic resin body fillers sold in automotive stores, with or without fiberglass reinforcement, are readily available. However, to make the filler flow readily in the cylinder, the filler should be heated so that it is fairly liquid. Don't use an open flame to make the filler liquid, but rather insert the container (a tin can will do) containing the filler into a bath of very hot water. Temporarily plug the top end of the PVC fitting where the central element protrudes and, with the cylinder initially held at an angle, pour in the filler from the coaxial cable end.

The use of a coaxial connector on the antenna was deliberately avoided. Simple coaxial connectors, when used in a harsh outdoor environment, will almost always eventually become the source of a problem. This is the same reason why screw-in elements, etc., are avoided. Of course, this all means that the antenna becomes a throw-away unit in case something should really damage it. But, the cost of materials involved to build another antenna is relatively low. As long as it does last, one can use the antenna with the confidence that none of the electrical or mechanical connections inside the antenna are likely to become corroded.

The 1/4λ cylinder need not be insulated from a metal mast as long as it is fastened to the mast at the bottom of the cylinder with the usual metal U-type mast clamps.

One can add parasitic or phased elements around the basic antenna if it is desired to obtain some directivity. Parasitic elements, of course, require no cable connection to the main element, but then one has to go through

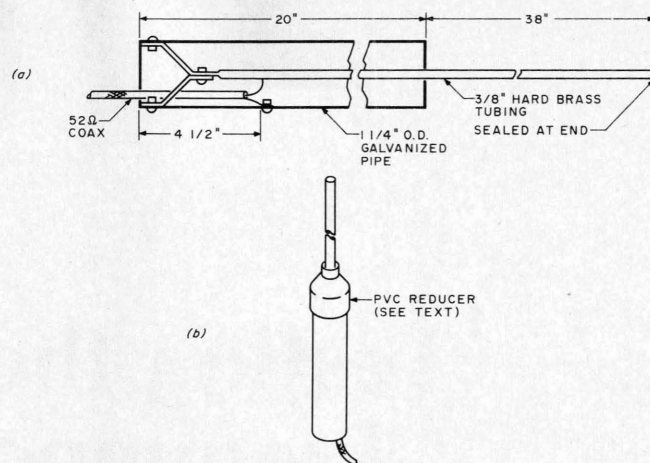


Fig. 2. (a) A cross-sectional view of the antenna giving dimensions centered in the 2 meter band. (b) A PVC pipe reducer used to insulate and center the vertical radiator at the top of the 1/4λ cylinder.

a careful process of seeing that the feedpoint of the main element is altered to compensate for the presence of the parasitic element. In spite of the increased cable cost, etc., if one really wants to keep the weather-ruggedness for a directive antenna in-

stallation paramount, it would be better to have two spaced antennas of the type described with separate feedlines. Then, the necessary phasing, matching, and switching can be done from the protected environment inside the shack. ■

BELDEN

Part Number	MHz	db/100 ft.	db/100 m
 9888 42¢/ft	50	1.2	3.9
	100	1.8	5.9
	200	2.6	8.5
	300	3.3	10.8
	400	3.8	12.5
 8214 26¢/ft.	50	1.2	3.9
	100	1.8	5.9
	200	2.6	8.5
	300	3.3	10.8
	400	3.8	12.5
 8237 23¢/ft	100	2.0	6.6
	200	3.0	9.8
	400	4.7	15.4
	900	7.8	25.6
 8267 27¢/ft	100	2.0	6.6
	200	3.0	9.8
	400	4.7	15.4
	900	7.8	25.6
 8448 17¢/ft	No. of Cond. — 8 AWG (in mm) — 6-22, (7x30), [1.76]; 2-18, (16x30), [1.19]		
 9405 28¢/ft	No. of Cond. — 8 AWG (in mm) — 2-16, (26x30), [1.52]; 6-18, (16x30), [1.17]		

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How To Succeed On 1296

—cat-food can 50-Watt amplifier

Amateur radio literature^{1,2,3} abounds with many excellent articles describing tube-type 2C39/7289 1296 MHz triplers and/or

amplifiers. Most of these require an inordinate amount of machine shop work not available to most amateurs. With the advent of readily-

available moderate-cost varactor triplers⁴ with power output after filtering in the 5-Watt range at 1296 MHz, it was believed there would be interest in a nominal 10 dB gain amplifier using a cavity that would enhance and increase $\frac{1}{4}$ meter DX activity during VHF contests.

Making the following amplifier requires only a circle cutter, sheet-metal shears, and a few socket punches. A small 18-inch Lafayette sheet-metal brake to bend up the .016" brass chassis would be a help, but it isn't necessary, as it may be bent up in an ordinary small vise.

Construction

The cavity bypass capacitor, part A (Fig. 2), top of cavity, part C (Fig. 3), and chassis, part E (Fig. 4), are cut from .016" sheet brass. The vertical cavity wall, part D (Fig. 3), is cut from a Purina 6-3/4 oz. cat-food (sardines) can or, alternatively, a Friskies Buffet Dixie Dinner 6½ oz. cat-food can (note: My cat recommends the former). Parts G, H, I, and J (Fig. 5) are press-fit tube contact rings made out of .010"

brass sheet. They are assembled in the following sequence:

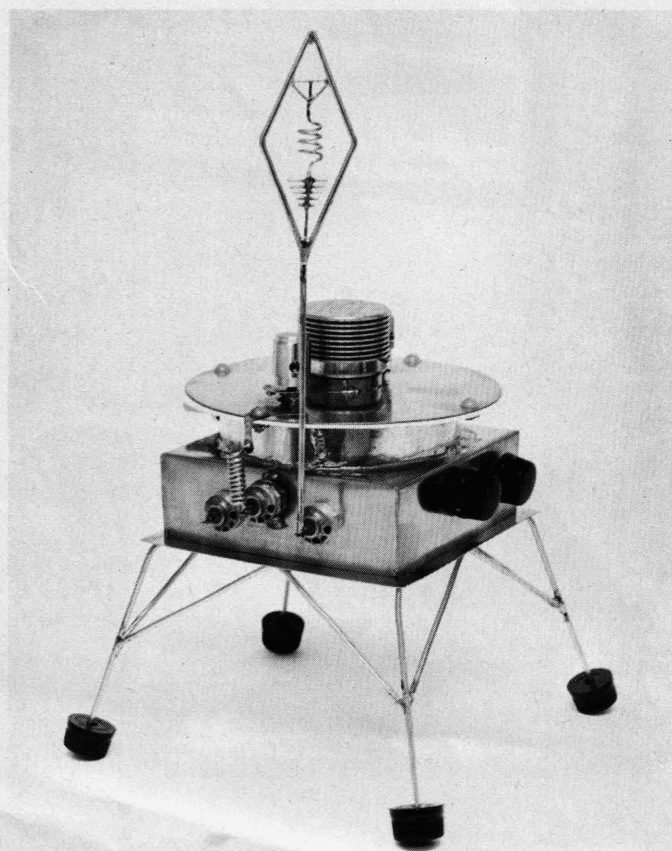
1. Wrap part G around the 2C39/7289 anode, and secure it with 2 pieces of #18 bare hookup wire twisted tightly with pliers. Make sure the ends of part G do not quite touch each other. Slip the tube with part G wired on just barely into part A so that the bottom of parts G and A are flush. Tack solder G to A about every quarter inch on the upper surface before smoothly flowing the solder around the full circumference. Use plenty of Nokorode soldering paste and an Ungar #4033 50-Watt soldering element to make the job easy. Remove the tube from part A. It should be a good tight press-fit, but, by slowly twisting as you pull, it should be readily removable.

2. Wrap part H around the tube's grid ring, and secure it with 2 or 3 wires, as above. Insert the tube into the chassis' center hole until part H is flush with the top of the chassis. Solder.

3. Wrap part I around the tube's bottom outer cathode ring and secure it with wire. Insert into part F all but 1/16" of part I, and solder.

4. Part J is made by wrapping around a 9/16" drill and squeezing with pliers until it makes a snug press into the inner filament ring in the tube base.

5. Center part A on top of part B on top of part C and drill the 4 outer holes each 9/64" diameter. Bolt the 3 parts together with conventional 1/4" 6-32 nuts and bolts. Solder the 4 nuts to the bottom of part C. Also drill the two 9/64" diameter holes for the 2 tuning capacitors, and disassemble. Run two ordinary 1/4" 6-32 nuts and bolts through the holes in part C, and solder the nuts to the top of part C. Drill 3/8" diameter holes through part A's tuning capacitor holes, and then reassemble parts A, B, and C with four 1/4" 6-32 nuts and bolts each.



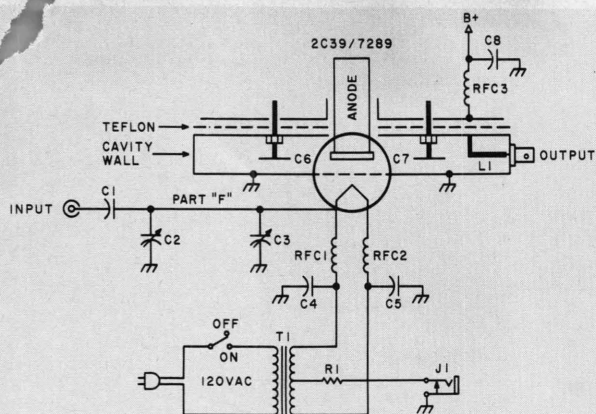


Fig. 1. Schematic.

6. With scissors cut a 3/4"-wide section of the cat-food can of your cat's choice. File it smooth, or, better yet, sand it smooth on a belt sander. With parts A, B, and C still bolted together, turn them upside down on a flat surface and tack solder part D to the bottom of part C before flowing the solder smoothly around the entire circumference.

7. Install and solder the output link and BNC connector as shown. Thread in the two tuning capacitors in the bottom of part C.

8. Install the tube in part G and A so that the top of part G is flush with the bottom of the protruding anode ring (tube is in as far as it will go). Carefully insert the tube into the chassis and part H until the bottom of the cavity is flush with the top of the chassis. There should be no radial or axial side loads on the tube if you have followed these steps in sequence — just a comfortable press-fit. Now, maintaining a gentle downward pressure on the tube, tack solder the cavity to the chassis before smooth flowing solder around the entire circumference.

9. Install the input BNC connector. Press-fit Parts F and I onto the cathode ring as far as they will go. The input blocking capacitor is modified by filing each side of a 500 pF ceramic disc cap until half of the leads have been filed away. Carefully tin each side with a small (25 Watt or

less) soldering iron. Solder in the 2 pi-net capacitors as shown. Solder the 500 pF modified disc cap to the end part F and to the inner conductor of the BNC connector. This is a fragile part and easily broken when removing and/or installing tubes. To further mechanically stabilize part F, take two burned-out 3/4"-long glass fuses and solder them to part F on the side opposite the two pi-net capacitors. They are structurally quite strong, good insulators, and they are free.

Operating

If you are like me, unable

to resist buying those one buck each 2C39 tubes at hamfests, you are most probably, also like me, throwing your money away. Of the first twelve tubes acquired this way, one was good; the others gave only marginal gain or were worthless. Jaro Electronics⁵ offers factory-new 2C39s or 7289s in quantities of 5 each that cannot be beat. I prefer the 7289 tubes with the more heat resistant ceramic seals.

It is best to tune up this amplifier at a reduced plate

voltage of, say, 300 volts. Use W1HDQ's favorite microwave 50-Ohm load, consisting of 100 feet of RG-58 coax. If you are an optimist, by all means terminate the coax with 10 each 1000-Ohm, 1-Watt resistors in parallel. Be assured that you will not burn them out.

Using a Bird ThrulineTM wattmeter with plate power off, and a 25-Watt full-scale 1.0-1.8 GHz slug, adjust the input pi-net capacitors for zero reflected power with approximately 5 Watts input. If

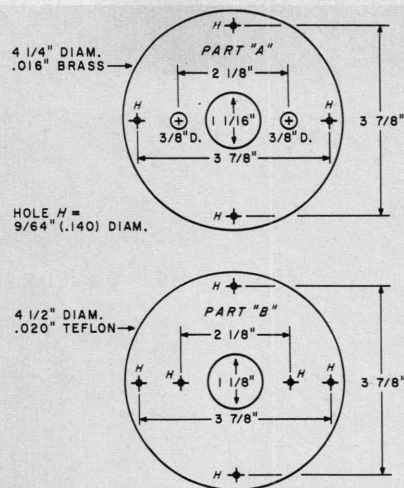
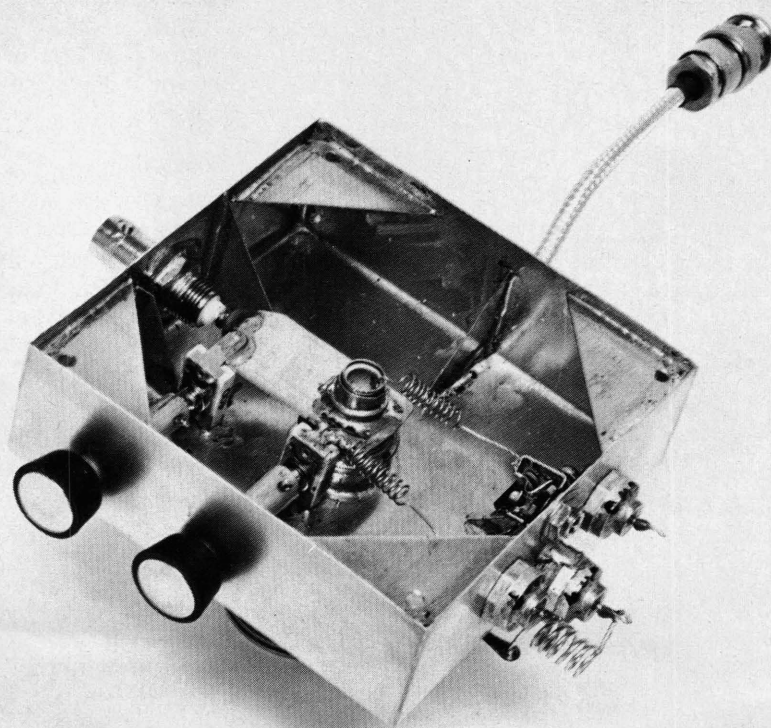


Fig. 2.



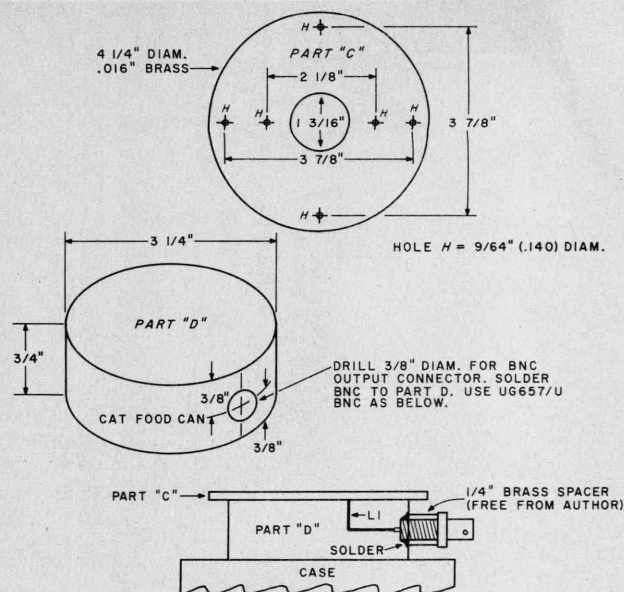
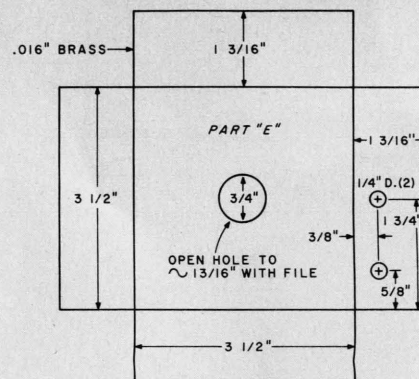
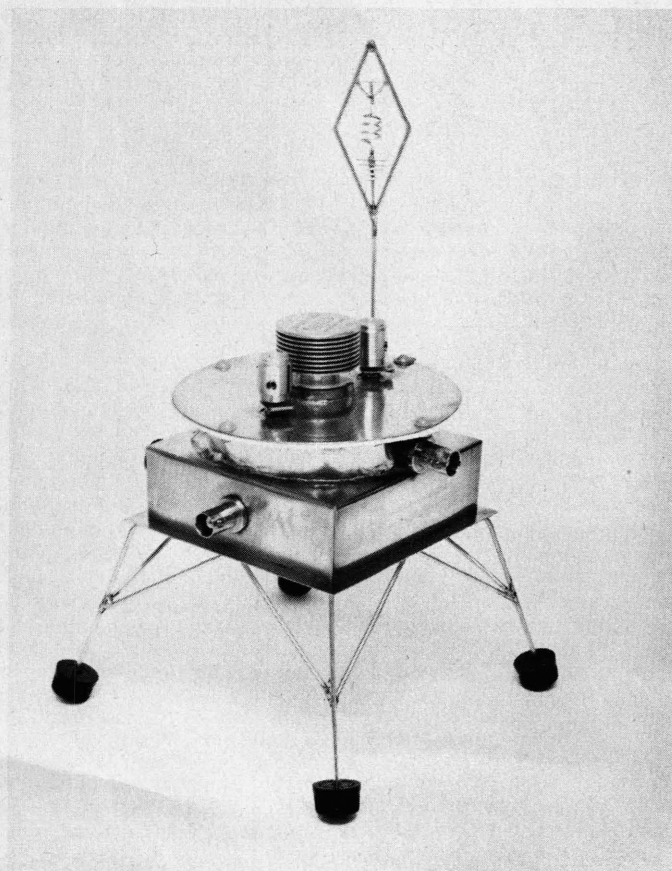


Fig. 3.

you are using a varactor tripler, you must have a good 1296 MHz filter between the tripler and amplifier input, otherwise your measurement will be totally meaningless as you will probably be measuring as much power at frequencies not on 1296 MHz as you are on 1296 MHz. With a milliammeter in series with

the cathode resistors, you should obtain 30 to 50 mA of grid current with the plate supply off. Remove the milliammeter.

With the Bird Thruline™ wattmeter in the cavity's output line and your dummy load attached, you should indicate about 1-Watt output, still with no voltage on the



NOTE: TUNING SHAFTS FOR C2 & C3 MADE AS FOLLOWS:

- ① NOTCH 3/8" DEEP WITH RAZOR BLADE.
- ② CUT .016" BRASS 5/16" X 1/2".
- ③ INSERT BRASS LONGWAYS INTO DOWEL NOTCH. WITH NO. 60 DRILL BIT DRILL HOLE THRU DOWEL & BRASS PIECE 2" 1/4" FROM END. INSERT SHORT PIECE NO. 18 WIRE THRU BOTH & SOLDER ENDS.
- ④ SOLDER ARCO 400 TUNING SCREW TO END OF BRASS PIECE "2".
- ⑤ YOU NOW HAVE A UNIVERSAL JOINT TUNING SHAFT FOR C2 & C3.

Fig. 4.

plate, when the cavity is properly tuned. Normal position of the tuning capacitors is about 1/8" from full IN. Though the aluminum knobs make a pretty photo, it is best to remove them and use 2 1/2" x 1/4" diameter wood dowels drilled to screw on the 6-32 tuning capacitors for adjustment when the plate voltage is ON, or you will be in for a "shocking" surprise.

With a new tube, you should obtain results as shown in Fig. 6 at voltages indicated.

Conclusions

If you are adept with a soldering iron and sheet-metal shears, it is much easier and quicker to build a sheet-brass cavity than to hog it out of brass ingots with a milling machine and lathe. A tin-plated cat-food can cavity shows no measurable difference in output or efficiency

when compared with the same tube in a similar silver-plated cavity. This is not to infer that a silver-plated cavity would be no better than one made out of kraft paper or Plexiglas™, just that the IR losses of cavity walls made out of cat-food cans is less than expected and not measurable with ordinary test equipment.

Should you wish to get on 1296 MHz and are either too lazy to roll your own or too affluent to wish to, try Spectrum International's filters, varactor triplers, receiving converters, and antennas. I have no association with this reputable firm in any form whatsoever, but I would like to see this band more populated to help preserve our precious amateur spectrum and to make available more VHF contest points, too.

Virtually every ham transmitter/amplifier article ends

Parts List

C1	500 pF (see text) or ceramic chip
C2, C3	Arco 400 (see Fig. 1)
C4, C5	500 pF feedthrough
C6, C7	see Figs. 1 and 5
C8	500 pF button bypass
L1	see Figs. 1 and 5
RFC1, RFC2	10 turns #26 1/8" diameter x 1" long
RFC3	10 turns #18 1/8" diameter x 1" long
R1	50 to 270 Ω, 1/2 Watt
T1	6.3 V c-t @ 1 Amp
J1	normally-closed mini jack

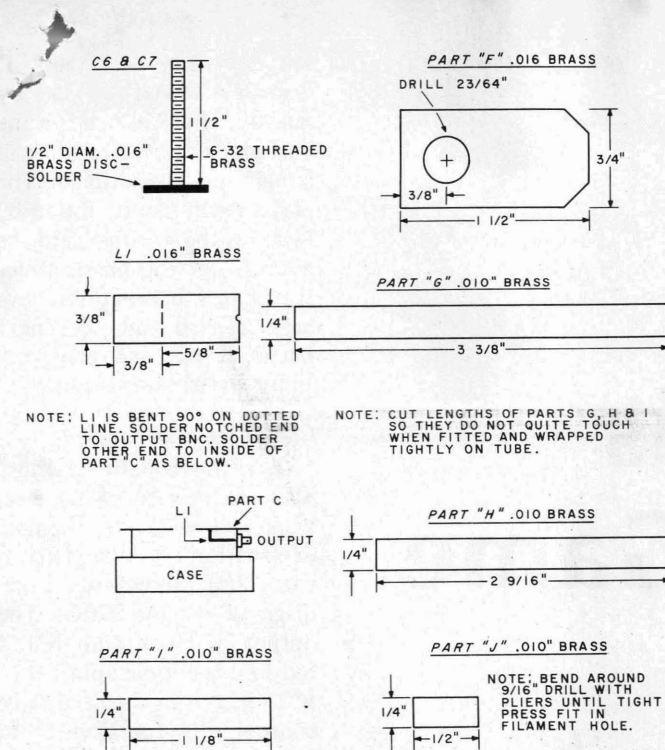


Fig. 5.

with the author's⁶ proud comment: "My first contact with the HBR11 was with ZE2ABS in Rhodesia." He called me while I was tuning

up with the much over-the-hill A, B, and C batteries connected and only a delta matched curtain rod in the basement for an antenna. I

will not disappoint readers who have read this far.

While tuning up this amplifier the first time on the air, I was pleasantly surprised by K2UOP's voice saying, "I hear you, I hear you!" Actually, I should stop right now, but Boy Scout honesty will out. K2UOP is a portable 4 living just 10 miles away, and his "I hear you" voice was on the telephone. We have worked with each other the last 3 VHF contests on 1296 MHz, and, though I am trying, I still have not burned out his receiving converter's first stage. ■

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2. Laakman, "Cavity Amplifier For 1296 Mc.," Jan, 1968, *QST* and *ARRL Handbook*, 1968 - 1972.
3. *R.S.G.B. VHF-UHF Manual*, pp. 6.48-6.52, 1969 edition.
4. Microwave Modules Ltd. models MMV 1296 and MMV 1296H - distributed in the U.S. by Spectrum International, Inc., Box 1084, Concord MA 01742.
5. Jaro Electronics, P.O. Box 414, Orlando FL 32802, in quantities of 5 each: 2C39 - \$10.00 each and 7289 - \$12.00 each.
6. Richardson, "High-Power VHF Triode Amplifiers," *QST*, July, 1959; "K.W. Amplifier For 6 and 2 Meters," *QST*, June, 1963.



EDITORIAL BY WAYNE GREEN

from page 8

much in the way of amateur radio and certainly not enough to be of any serious help in emergencies. Few of these countries even have a rough idea of why amateur radio can be of any benefit to them...

Very few countries are going to voluntarily give up radio frequencies, which they feel they need, for something which is of no value to them—hams. They know that even if they get more radio channels than they need for their own immediate use, they will be able to rent these frequencies out to commercial interests... and the going rate is about \$10 million per channel. The recent ITU maritime conference made this fact of life clear and inescapable.

Since amateur radio could be an enormous asset to any third world country, it is a

shame that no coordinated effort whatever is being made to make these countries aware of what they are missing and how they could go about getting these benefits.

SALTING THE WOUND

As bad as the Boston situation was after the blizzard of '78, this would be nothing compared to what would be needed in communications should there be a nuclear war. Yes, I know this is an old harp and we aren't really worrying about that any more... war is not an alternative any longer... etc.

One of the reasons for our almost impenetrable complacency is the SALT agreement which essentially placed the populations of the major cities of the U.S. and U.S.S.R. in the role of hostages. Both sides agreed to cease efforts to protect the city populations as a deterrent to massive attack and massive counterattack.

As any of you who are involved with Civil Defense know all too well, the U.S. has not only lived up to this agreement, but has bent over backwards not to make any effort to protect city populations. We have no shelters, no food storage, no drills, etc.

Contrast this, if you will, with the Soviet Union, where virtually every factory has a built-in shelter, equipped to feed and protect the workers until the radiation from a nuclear bombing would be low enough to reap a new harvest. The Soviet cities are riddled and ringed with shelters, built deep enough to survive just about anything. They have air filters and extended-term living facilities to await fallout contamination.

The ugly fact is that we've been taken to the cleaners again by honoring an agreement with Russia. This is of no serious consequence as long as no international crisis arises where we feel we must rattle our nukes. As long as we are content to let Russia, with the help of satellite Cuba, take over one country in Africa after another, we have no worry. They grabbed Angola and have the top hand in a half dozen other African countries. Now they are well along with taking

over Somalia, Ethiopia, Djibouti, and environs.

In personal relations, in business, and in politics, the weak and indecisive lose out... and how else could you characterize the U.S. in recent years?

Okay, what does all this mean to amateur radio? Unless Civil Defense is brought back to life and changes beyond imagining are made, any nuclear exchange would inconvenience about 80% of the U.S. population, with about 50% inconvenienced to the point of death. In a recent statement, the Chairman of our Joint Chiefs of Staff estimated that we would lose ten people for every Russian killed—thus we would lose about half our people in a nuclear exchange while Russia would lose about 4% of theirs... half what they lost during WW II.

With a situation like that, the only means of communications that would be usable would be amateur radio. Telephones would be out since they are controlled from the cities... and cities would be wiped out. There is no other radio communications service which covers local and long ranges... just amateur radio.

Continued on page 38

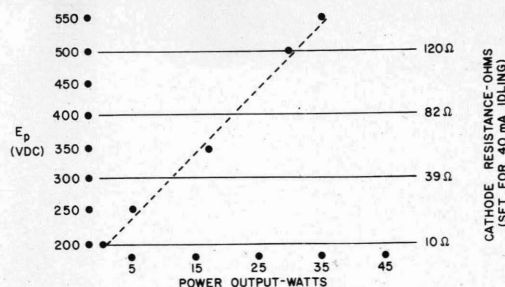
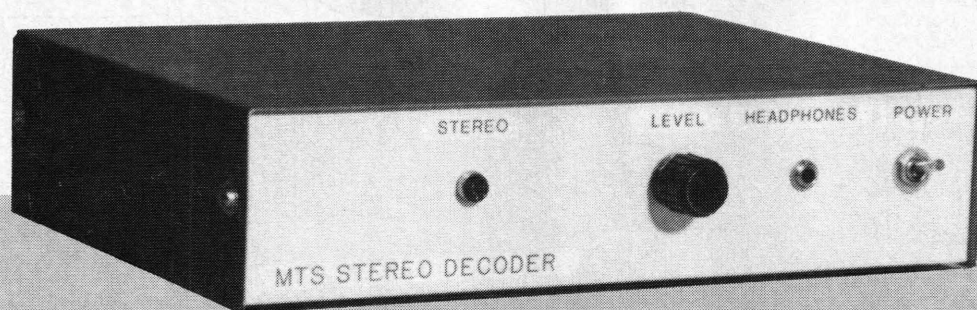


Fig. 6. Input held at 4 Watts.



TOD T. TEMPLIN

STEREO TV DECODER

Build your own high-fidelity MTS decoder for the finest in TV enjoyment.

A RECENT SURVEY OF TELEVISION STATIONS across the U. S. and Canada reveals that more than 250 stations are now transmitting MTS stereo TV sound. So chances are good that at least one station in your area is transmitting stereo audio right now. You might think that you need a stereo TV or VCR to enjoy MTS, but consider this: For about \$50 (for all new parts), you can build our add-on converter, which will work with virtually any TV or VCR. All components are readily available, and we've designed a PC board, which simplifies construction greatly. The circuit may be aligned by ear, although using an oscilloscope will give more precise results.

Background

To understand how we can enjoy MTS sound, let's look back to when color-TV standards were formed. In 1953 the NTSC (National Television Systems Committee) defined the standards for color-TV broadcasting that are now used in the U. S., Canada, Mexico, and Japan.

In the NTSC system, 6 MHz is allocated for each television channel, as shown in Fig. 1. Video information is transmitted on an amplitude-modulated carrier that extends about 4.2 MHz above the visual carrier. Mono audio is transmitted on a frequency-modulated carrier 4.5 MHz above the video carrier, with 100% modulation causing a 25-kHz deviation of that carrier. So a fully modulated mono signal causes the carrier to vary between 4.475 and 4.525 MHz around the carrier.

By subtracting 4.2 MHz (top of video) from 4.475 MHz (bottom of audio), we find that there is 275 kHz of unused spectrum. That space was originally allocated as a guard band by the NTSC. The reason the guard band was necessary was that the tube-based circuits of that era were less

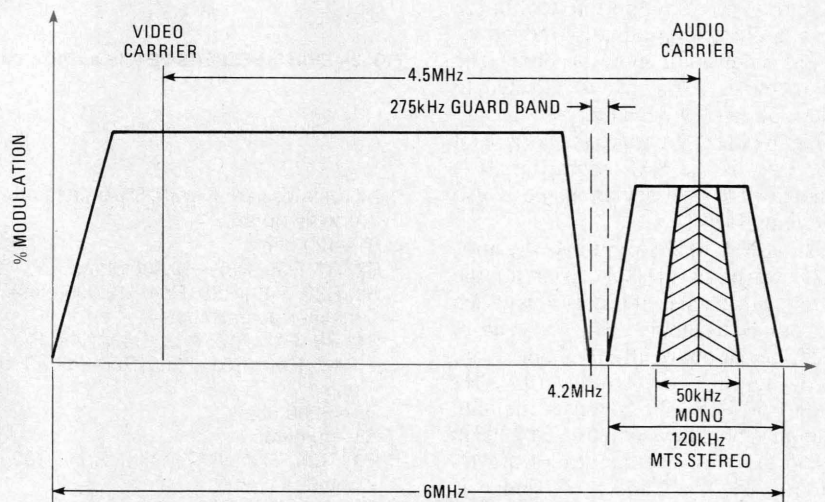


FIG. 1—STEREO-TV AUDIO requires about twice the bandwidth of a mono signal.

capable of keeping the audio and the video portions of the signal separate than modern solid-state circuits. It is that 275-kHz gap that allows us to have MTS sound today.

On March 29, 1984, the BTSC (the Broadcast Television Standards Committee, which is the present-day equivalent of the NTSC), proposed guidelines to the FCC (in BC docket 21323) for TV stations using the BTSC system of multichannel sound transmission. That docket contains general technical rules governing the use of the television audio baseband for use in the transmission of stereo television sound, as well as a second-language channel (SAP, for Second Audio Program), and a professional channel. (The alternate services were discussed in "Stereo Audio for TV," *Radio-Electronics*, February and March 1985, and in "Stereo TV Decoder," in the March 1986 issue of *Radio-Electronics*.—Editor)

As in the NTSC system, the baseband mono audio signal (which is the equivalent of the L+R stereo signal) has a bandwidth of about 15 kHz. It is transmitted with 75 μ s of pre-emphasis, and has a maximum deviation of 25 kHz.

At 15.734 kHz is the BTSC pilot tone. The pilot is locked to horizontal sync, and it is used to identify the signal as a BTSC transmission, thus informing the television receiver to switch from mono to stereo reception. The pilot has a 5-kHz deviation.

Then comes the stereo difference signal (L-R). It is amplitude modulated on a 31.468 kHz subcarrier, producing a double-sideband suppressed-carrier signal that spans about 30 kHz. That subcarrier frequency was chosen because it is exactly twice the NTSC horizontal sweep frequency, and is, therefore, easily synchronized during both transmission and reception.

NOISE REDUCTION

THE STEREO DECODER DESCRIBED IN THIS article doesn't use a true dbx decoder. When we first decided to build an MTS decoder, we contacted the engineers at dbx Corporation in an attempt to obtain engineering samples of their decoder IC's. As you may know, however, dbx Corporation does not sell those IC's to unlicensed persons or companies, and that includes hobbyists. We were discouraged, but decided to go ahead and build a converter without the dbx IC's, and see just how well it could be done.

The decoder presented here is the result of that effort, and we believe that it performs as well as many commercial units. In addition, none of the electronic components used are difficult to obtain. Also, due to a very flexible design, you can interface the decoder to almost any TV or VCR and obtain very good results. **R-E**

The L-R signal is also compressed by a complex noise-reduction technique known as dbx television noise reduction. (See the sidebar for more on dbx.) The level of the L-R signal is adjusted to produce 50 kHz of deviation.

At 78.67 kHz (five times the horizontal sweep rate) is the SAP subcarrier. It is limited to 10-kHz of deviation and is also dbx compressed.

Last, at 102.3 kHz (6.5 times the horizontal sweep), is the subcarrier for the professional channel. It is not compressed and is limited to about 3-kHz of deviation.

If the deviations of all sub-channels are added together, the total is 98 kHz (25 + 5 + 50 + 15 + 3). However, the total deviation is not allowed to exceed 73 kHz (50 + 15 + 3), because the sum of the deviations of the L + R and L - R signals is limited to 50 kHz. Although that total is greater than the deviation of a plain mono transmission, it fits into the guard band with room to spare.

If you're familiar with the stereo system used for FM radio transmissions, you'll notice that the stereo portion of the BTSC system is essentially the same as that used in FM radio, disregarding the SAP and professional channels. In fact, the main differences are the slightly different frequencies of the pilot and the L-R subcarriers. We can take advantage of those similarities by using an IC that is normally used to decode FM radio signals. Doing so simplifies our design and reduces costs considerably.

The circuit

A block diagram of the stereo-TV decoder is shown in Fig. 2. It shows the overall relationships between the separate sections of the circuit; Figures 3-6 show the details of each subsection.

Let's start with the decoder section (shown in Fig. 3). It centers around IC1, a

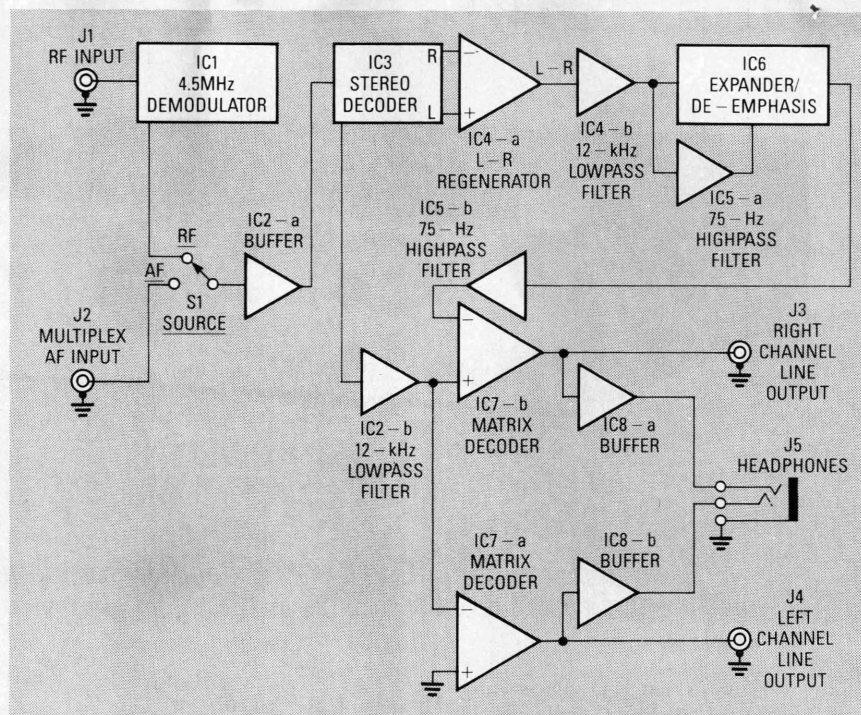


FIG. 2—EIGHT INEXPENSIVE IC'S are all it takes to provide a high-quality MTS decoder.

PARTS LIST

All resistors are 1/4-watt, 5% unless otherwise noted.

R1—120 ohms
R2, R7, R35, R37—10,000 ohms
R3, R23, R49, R53, R54—10,000 ohms, trimmer potentiometer
R4, R6, R11, R12, R42, R43, R44, R46, R48, R50, R51, R59, R60—100,000 ohms
R5—2200 ohms
R8—10 ohms
R9, R24, R31, R57, R58, R63—1000 ohms
R10, R16, R17, R28—3300 ohms
R13—330,000 ohms
R14, R15, R21, R62—4700 ohms
R18—12,000 ohms
R19—25,000 ohms, trimmer potentiometer
R20—4300 ohms
R22, R27—5100 ohms
R25—5,000 ohms, trimmer potentiometer
R26—1500 ohms
R29—30,000 ohms
R30—18,000 ohms
R32, R33, R39, R40—20,000 ohms
R34, R41, R55, R56—39,000 ohms
R36, R38—22,000 ohms
R45—68,000 ohms
R47—470,000 ohms
R52—100,000 ohms, dual-gang potentiometer
R61—330 ohms

Capacitors

C1, C4, C13, C32—0.01 μ F, ceramic disk
C2, C9, C19—470 pF, ceramic disk
C3, C14—0.05 μ F, ceramic disk
C5—5-60 pF, trimmer
C6—10 pF, ceramic disk

C7, C8, C10, C11, C27, C38, C47—1 μ F, 50 volts, electrolytic
C12, C23, C25—0.0022 μ F, ceramic disk
C15, C30, C34—C37—0.22 μ F, ceramic disk
C16, C17—0.47 μ F, ceramic disk
C18—0.0047 μ F, ceramic disk
C20, C21—0.0015 μ F, ceramic disk
C22, C24—0.0039 μ F, ceramic disk
C26, C29—0.015 μ F, ceramic disk
C28, C31, C39—C46—10 μ F, 50 volts, electrolytic
C33, C50—C53—2.2 μ F, 50 volts, electrolytic
C48—2200 μ F, 50 volts, electrolytic
C49—470 μ F, 50 volts, electrolytic

Semiconductors

IC1—MC1358 stereo demodulator
IC2, IC4, IC5, IC7, IC8—LM358 dual op-amp
IC3—LM1800 stereo decoder
IC6—NE570 compander
D1, D1—1N4002 rectifier diode
LED1, LED2—standard
Q1, Q3—2N3904 NPN transistor
Q2—2N3906 PNP transistor
Q4—2N2222 NPN transistor

Other components

F1—1/4-amp, 250-volt fuse
J1—J4—RCA phono jack
J5—stereo headphone jack
L1—33 μ H S1—SPDT toggle switch
S2—SPST toggle switch
T1—10.7 MHz IF transformer
T2—25-volt CT power transformer

Note: A drilled, etched, and plated PC board is available from Tod. T. Templin, 5329 N. Navajo Ave., Glendale, WI 53217 for \$9.00.

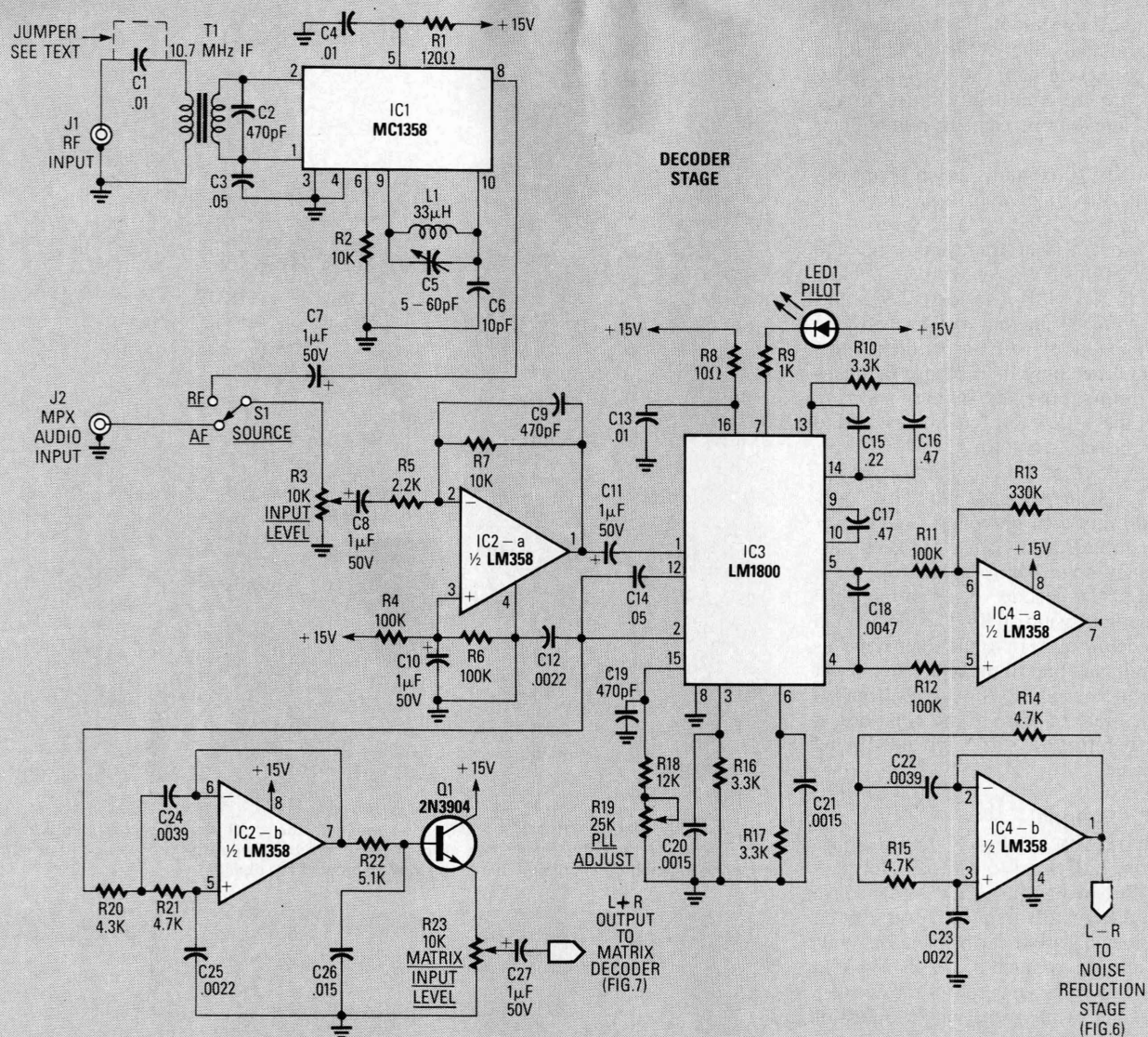


FIG. 3—THE DECODER STAGE converts the multiplexed audio signal into L + R and L - R signals.

standard 4.5-MHz audio demodulator that is used in many television receivers. The circuit is more or less what you find in the databook. The major exception is that the standard de-emphasis capacitor has been eliminated in order to ensure that the L - R signal is presented to the decoder. If present, the capacitor would roll off the high-frequency L - R signal.

The output of IC1 is routed to S1, which allows you to choose between the internally demodulated signal and an externally demodulated one. Buffer amplifier IC2-a then provides a low-impedance source for driving IC3, an LM1800 stereo demodulator. As with IC1, IC3 is used in a conventional manner. Our circuit differs from the cookbook circuit, however, in that the component values associated with the phase-locked loop have been altered so that the loop will lock on the 15.734-kHz MTS pilot rather than on the 19-kHz FM-radio pilot.

When IC3 is locked on a stereo signal,

the outputs presented at pins 4 and 5 are the discrete left- and right-channel signals, respectively. In order to provide noise reduction to the L - R signal, we must re-combine the discrete outputs into sum and difference signals. Op-amp IC4-a is used to regenerate the L - R signal. It is wired as a difference amplifier, wherein the inputs are summed together (+ L - R). Capacitor C18 bridges the left- and right-channel outputs of the demodulator. Although it decreases high-frequency separation slightly, it also reduces high-frequency distortion. After building the circuit, you may want to compare sound output with and without C18.

The L + R signal is taken from the LM1800 at pin 2, where it appears conveniently at the output of an internal buffer amplifier.

The raw L - R signal is applied to IC4-b, a 12-kHz lowpass filter. The L + R signal is also fed through a 12-kHz lowpass filter in order to keep the phase shift un-

dergone by both signals equal. If only one were filtered, there would be a loss of high-frequency separation when the left and right channel signals were recovered.

Next, as shown in Fig. 4, the L - R signal is fed to Q2. That transistor has three functions. It allows us to add a level control to the L - R signal path; it provides a low source impedance for driving the following circuits; and it inverts the signal 180°. (Think of the signal at the collector of Q2 as -(L - R)). Inversion is necessary to compensate for the 180° inversion in the compander.

Next comes the expander stage; this is where we would use a dbx decoder if we could get one (see sidebar). At the collector of Q2 is a 75-μs de-emphasis network (R27 and C29) that functions just like the network associated with Q1 (in Fig. 3). Note that Q2 feeds both Q3 and IC5-a, a -12 db per octave highpass filter. The output of that filter drives the rectifier input of IC6, an NE570.

The NE570 is a versatile compander. We'll use it as a simple 2:1 expander. The 75-Hz highpass filter at the rectifier input helps to prevent hum, 60-Hz sync buzz, and other low-frequency noise in the L-R signal from causing pumping or breathing.

The NE570 contains an on-board op-amp; its inverting input is available directly at pin 5, and via a 20K series resistor at pin 6. That's a convenient place to implement the 390- μ s fixed de-emphasis network. The 18K resistor (R30) combines with the internal resistor and C32 (0.01 μ F) to form a first-order filter with a 390- μ s time constant. Because the internal op-amp operates in the inverting mode, the $-(L-R)$ signal is restored to the proper $(L-R)$ form.

The output of the expander drives another 75-Hz highpass filter, but this one is a third-order type providing -18 db per octave rolloff. It too is used to keep low-frequency noise from showing up at the output of the decoder. Keep in mind the fact that television audio does not extend much below 50 Hz, so the filter removes no significant part of the audio signal. At this point the $(L-R)$ signal has been restored, more or less, to the condition it was in before it was dbx companded at the transmitter.

The L + R signal

Referring back to Fig. 3, the L + R signal from IC3 is fed to a 12-kHz lowpass filter, IC2-b, with a -12 dB per octave slope. That cutoff frequency was chosen in a somewhat arbitrary manner. We wanted to remove as much of the 15.734-kHz pilot signal from the output of the decoder as possible, while preserving as much of the desired high-frequency audio as possible. So we settled on 12 kHz as a good compromise.

The output of the highpass filter is applied to a 75- μ s de-emphasis network (R22 and C26). The L + R audio signal is now restored properly. We feed it through Q1, which is wired as an emitter follower to provide a high load impedance for the de-emphasis network and a low source impedance for level control R23. Next the L + R signal is fed to the matrix decoder, shown in Fig. 5.

Left and right recovery

Op-amps IC7-a and IC7-b are used to recover the individual channels. First, IC7-b is configured as unity-gain difference amplifier. The $(L+R)$ is applied to its inverting input, and the $(L-R)$ signal is applied to the non-inverting input. Therefore the output of IC7-b may be expressed as $-(L+R) + (L-R) = -L + L - R - R = -2R$. Similarly, IC7-a is configured as a mixing inverting amplifier. Here, however, both sum and difference signals are applied to the inverting input. So the output of IC7-a is $-(L+R)$

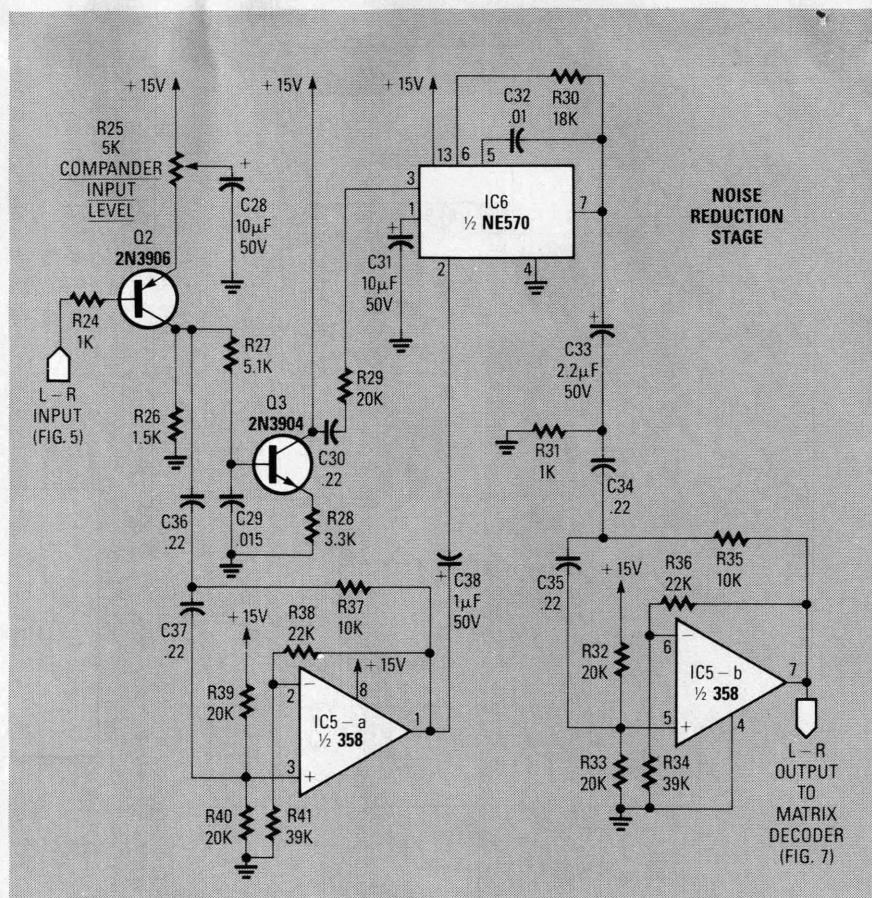


FIG. 4—THE NOISE-REDUCTION STAGE de-compands the L-R signal, and emulates dbx-style processing. As described elsewhere in this article (see box), true dbx processing is not currently possible in a home-built circuit due to the inavailability of the dbx IC's.

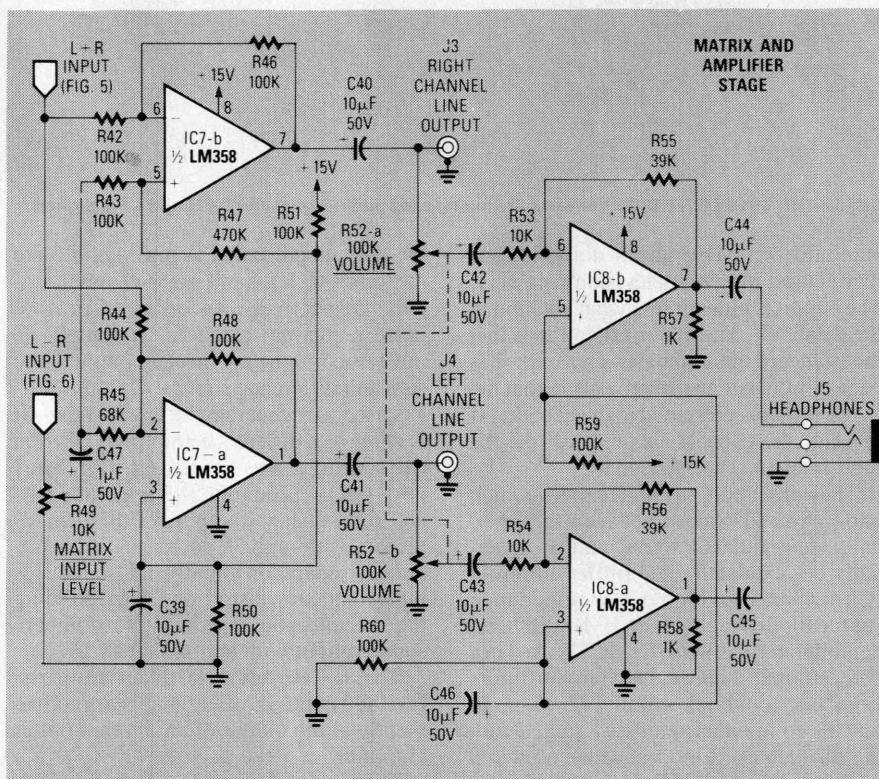


FIG. 5—THE MATRIX STAGE separates the L + R and L - R signals into the left- and right-channel components. Op-amp IC8 and associated components provide an optional headphone output. If you do not wish to drive a pair of headphones, or plan to use your amplifier's headphone jack for that purpose, all components to the right of jacks J3 and J4 can be deleted.

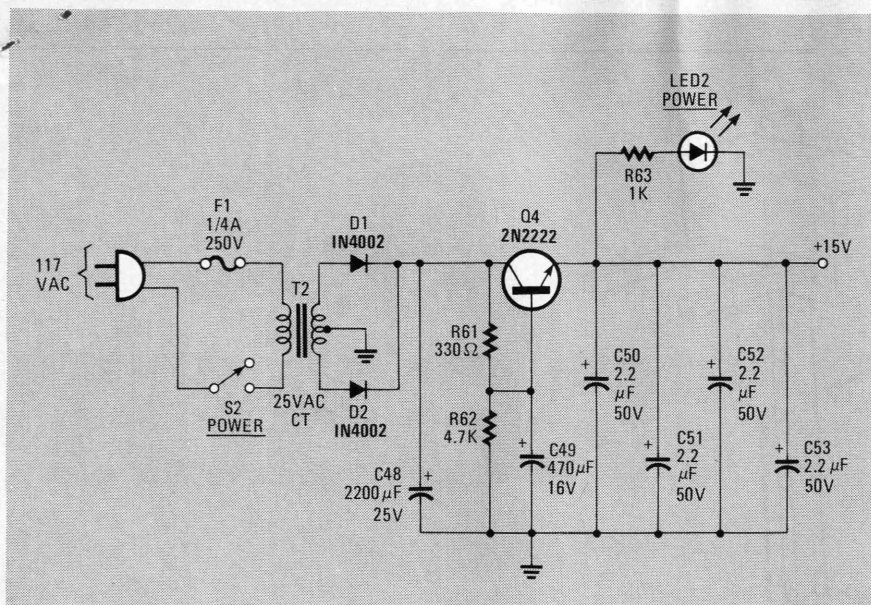


FIG. 6—THE UNREGULATED POWER SUPPLY shown here provides extremely low ripple for the MTS decoder.

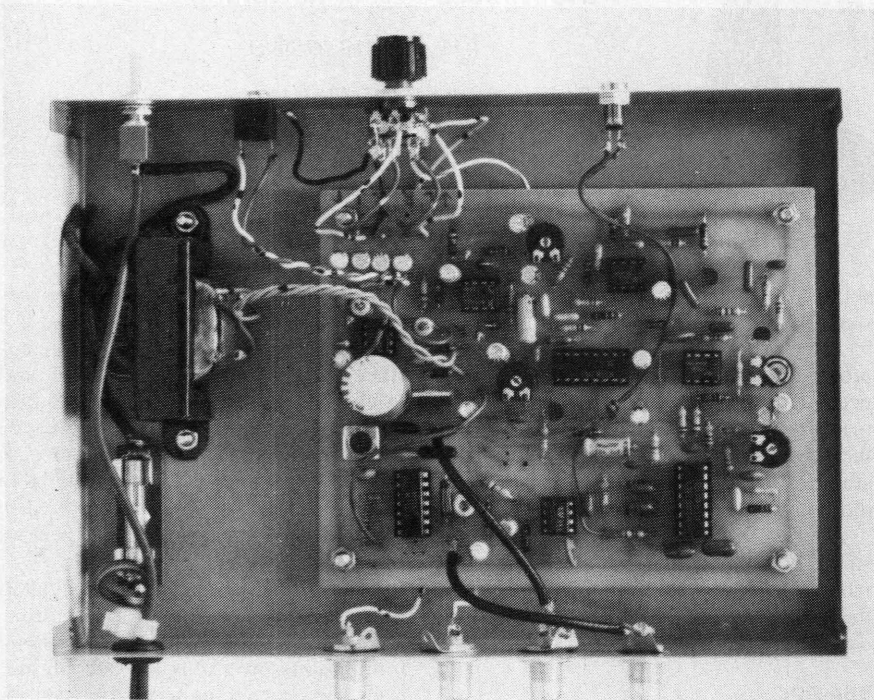


FIG. 7—THE COMPLETED STEREO DECODER BOARD. Next time, we'll show you how to build the circuit shown here.



AN OUTBOARD TUNER lets you use the circuit with a TV that lacks audio outputs. Next time, we'll see many other ways of using the circuit.

$-(L-R) = -L - R - L + R = -2L$. Because both channels have been inverted, the stereo relationship is preserved.

The two op-amps in IC8 provide an additional stage of amplification to drive a pair of stereo headphones. If you don't plan to use headphones, or if you are content to use only your stereo's headphone jack, all components to the right of line-output jacks J3 and J4 may be deleted.

The schematic of the decoder's power supply is shown in Fig. 6. It provides an unregulated 15-volt DC output. Transistor Q4 is used as a capacitance multiplier, to

COMPENSATION

THE MARCH 1985 ISSUE OF **RADIO-ELECTRONICS** has a good description of the dbx system, but we'll summarize the salient features here. Keep in mind the fact that dbx operates only on the stereo difference signal ($L-R$).

- The signal is compressed at transmission by a fixed ratio of 2 to 1.
- The signal is pre-emphasized by a combination of 75- μ s and 390 μ s networks.
- The signal is spectrally companded by a variable ratio that depends on broadband frequency balance and signal level.

Of those three functions, spectral companding is the most difficult to compensate for. We include de-compression circuits and the proper de-emphasis networks, but we decided not to include spectral de-companding in our decoder, based on the following rationale.

Spectral companding's primary function is to mask high-frequency noise when the signal is composed primarily of low frequencies at relatively low levels. It does so by adding a variable amount of high-frequency pre-emphasis at the proper times. If the signal contains relatively high signal levels across the entire audio spectrum, little spectral companding is performed. Fortunately, in the real world of television broadcasting, high-level signals that extend across the entire audio spectrum are fairly common, so little dbx companding actually is performed.

All television stations use sophisticated audio processing devices to boost the audio level during quiet program material, and to limit the level during loud material (like commercials). Those devices generally divide the spectrum into three bands, and each band is independently monitored by the processor to ensure that the levels in each band remain relatively high.

The end result is that overall modulation remains high across the entire audio spectrum for most types of program material. Therefore, the dbx circuitry would do little spectral companding, so we made no attempt to compensate. **R-E**

provide high ripple reduction. The four 2.2- μ F capacitors (C50-C53) are distributed on the PC board (which we'll show next time) to keep the impedance of the power-supply rails low. That's important to minimize crosstalk between different sections of the unit.

As shown in Fig. 7, most of the circuitry we've described mounts on a single PC board. Unfortunately, we've run out of space for this month. When we continue we will show you how to build the circuit, as well as several methods of connecting the unit to a TV or VCR. At that time, the PC pattern will be provided. If you wish to get a head start, and are planning to purchase a pre-etched board, you can order one from the source provided in the Parts List. **R-E**

Q & A

CONDUCTED BY MICHAEL A. COVINGTON, N4TM1

Audio Induction Coil

Q I am 90 years old and have been reading *Gernsback electronics magazines* for many years. I once learned from your magazine how to hook up a Morse code key to a Model T spark coil to make a radio transmitter. It worked then, but today it would mess up a lot of people's TV reception.

My problem now is different. I need a schematic for a device to take the signal from a telephone circuit, amplify it, and feed it into a loop around a room. This should set up a magnetic field that can be detected by hearing aids with T coils. I know it has been done before. If you can be of any help on this, I and many of my senior friends would appreciate it.—H. B. A., Bay Village, OH

A What you describe is called audio induction. A large coil around the room and a small coil on the pickup unit form a big transformer, transferring the audio signal from one to the other magnetically. This isn't radio; little or no electromagnetic radiation is produced. (The coil is a monstrously inefficient antenna at audio frequencies.) Instead, what you're doing is exactly what goes on in transformers.

Some hearing aids use induction coils to pick up the signal from telephone receivers—it's more reliable than picking up the sound. Not having one of these hearing aids handy, we're not sure how sensitive they are, but Fig. 1 shows an audio induction circuit we've experimented with. The big coil has a resis-

tance of about 16 ohms and is fed with a few watts of audio, just as if it were a speaker. The small coil picks up a tiny audio-frequency signal, which is stepped up by the matching transformer and then fed to the microphone input of an amplifier. Pickup coils with substantially more turns may not need the transformer.

We trust you can adapt this circuit to your needs; you'll be using a hearing aid with an induction pickup in place of the pickup coil and amplifier. Others will find it useful as a way to transmit audio without wires to listeners who are free to move around and may even be on the other side of a wall.

Remote Control Repeater

Q I'm sitting in my living room in my favorite chair trying to find the right angle at which to bounce the infrared remote control signal off the ball mirror and change the CD. What I need is a remote control extender. I tried to use the RadioShack IR receiver to modulate an IR emitter but I had no luck. What am I doing wrong?—T. P., St. Albert, Alberta, Canada

A The solution to the mystery is that the IR signal is chopped (turned on and off) at about 40 kHz to distinguish it from ambient IR light. The receiver module supplies un-chopped output, so you have to chop it again if you want to use it to control an IR-emitting LED. We published a circuit that does this in

the December, 1995 installment of this column.

Sick Ohmmeter

Q I recently dusted off my Simpson 260-5 (circa 1965) volt-ohm-milliammeter (VOM) and I'm having some trouble with the $R \times 1$ scale readings. Fresh batteries and clean battery connections allow a full-scale 0 reading when I short the test leads, but the readings seem to be high by a factor of 4: 50-ohm resistors read 200 ohms, and so forth. I verified the test resistors with a DMM, VTVM, and another VOM. They all concur—the Simpson is off. The $R \times 100$ and $R \times 10K$ scales read normally, as do all voltage and current scales.

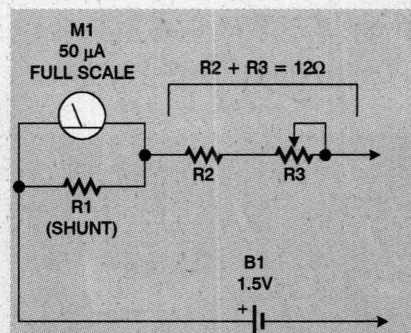


FIG. 2—THIS IS THE BASIC CIRCUIT of an ohmmeter that reads 12 ohms at half scale. The value of R1 is determined by the resistance of M1, the meter movement.

If anyone recognizes this anomaly, or has service experience with the 260, I would greatly appreciate any suggestions. It's a great meter that's worth saving. Also, I'd like to get a schematic and owner's manual for it; any ideas on where or how? Thanks.—J.A., Stony Point, NY

A Your second question is easy: Simpson Electric Company is still in business and still making the 260, although they're now up to model 260-8 instead of 260-5. You can reach them at 8853 Dundee Ave., Elgin, IL 60120, Tel: 847-

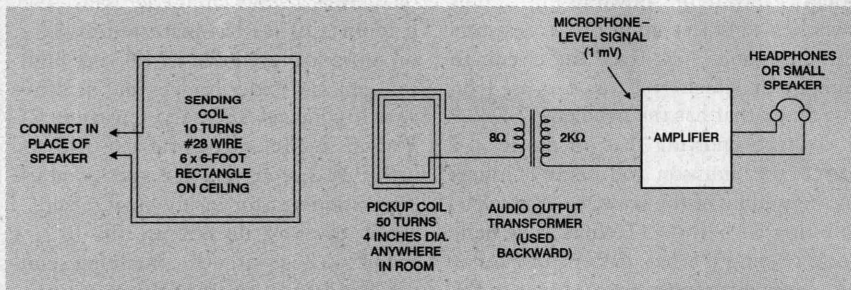


FIG. 1—AN INDUCTION LOOP transmits audio a few feet without wires. Experiment with different pickup coils; some might not need a transformer.

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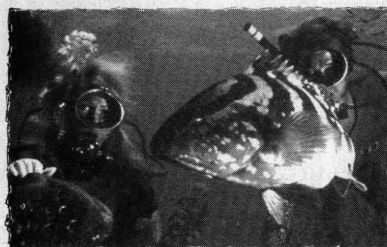
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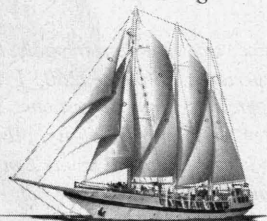


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HOW TO GET INFORMATION ABOUT ELECTRONICS

On the Internet: See our Web site at <http://www.gernsback.com> for information and files relating to our magazines (**Electronics Now** and **Popular Electronics**) and links to other useful sites.

To discuss electronics with your fellow enthusiasts, visit the newsgroups sci.electronics.repair, sci.electronics.components, sci.electronics.design, and rec.radio.amateur.homebrew. "For sale" messages are permitted only in rec.radio.swap and misc.industry.electronics.marketplace.

Many electronic component manufacturers have Web pages; see the directory at <http://www.hitex.com/chipdir/>, or try addresses such as <http://www.ti.com> and <http://www.motorola.com> (substituting any company's name or abbreviation as appropriate). Many IC data sheets can be viewed online. Extensive information about how to repair consumer electronic devices and computers can be found at www.repairfaq.org

Books: Several good introductory electronics books are available at RadioShack, including one on building power supplies.

An excellent general electronics textbook is *The Art of Electronics*, by Paul Horowitz and Winfield Hill, available from the publisher (Cambridge University Press, 1-800-872-7423) or on special order through any bookstore. Its 1125 pages are full of information on how to build working circuits, with a minimum of mathematics.

Also indispensable is *The ARRL Handbook for Radio Amateurs*, comprising over 1000 pages of theory, radio circuits, and ready-to-build projects, available from the American Radio Relay League, Newington, CT 06111, and from ham-radio equipment dealers.

Copies of past articles: Copies of past articles in **Electronics Now** and **Popular Electronics** (post 1994 only) are available from our Claggg, Inc., Reprint Department, P.O. Box 4099, Farmingdale, NY 11735; Tel: 516-293-3751.

Electronics Now and many other magazines are indexed in the *Reader's Guide to Periodical Literature*, available at your public library. Copies of articles in other magazines can be obtained through your public library's interlibrary loan service; expect to pay about 30 cents a page.

Service manuals: Manuals for radios, TVs, VCRs, audio equipment, and some computers are available from Howard W. Sams & Co., Indianapolis, IN 46214 (800-428-7267). The free Sams catalog also lists addresses of manufacturers and parts dealers. Even if an item isn't listed in the catalog, it pays to call Sams; they may have a schematic on file which they can copy for you.

Manuals for older test equipment and ham radio gear are available from Hi Manuals, PO Box 802, Council Bluffs, IA 51502, and Manuals Plus, PO Box 549, Tooele, UT 84074.

Replacement semiconductors: Replacement transistors, ICs, and other semiconductors, marketed by Philips ECG, NTE, and Thomson (SK), are available through most parts dealers (including RadioShack on special order). The ECG, NTE, and SK lines contain a few hundred parts that substitute for many thousands of others; a directory (supplied as a large book and on diskette) tells you which one to use. NTE numbers usually match ECG; SK numbers are different.

Remember that the "2S" in a Japanese type number is usually omitted; a transistor marked D945 is actually a 2SD945.

Hamfests (swap meets) and local organizations: These can be located by writing to the American Radio Relay League, Newington, CT 06111; (<http://www.arrl.org>). A hamfest is an excellent place to pick up used test equipment, older parts, and other items at bargain prices, as well as to meet your fellow electronics enthusiasts—both amateur and professional.

697-2260, Web: www.simpsonelectric.com. Many technicians prefer a traditional VOM because it can't be damaged by static electricity.

We don't have the schematic ourselves, nor is it necessary; the ohmmeter portion of a classic VOM always has roughly the circuit shown in Fig. 2, plus switches to select appropriate resistors for each range. Just trace the circuit to figure out which resistors are used on the range that has the problem.

In fact, knowing that on the $R \times 1$ range, the Simpson 260 reads 12 ohms in the middle of the scale, we can calculate that $R_2 + R_3 = 12$ ohms. The half-scale reading is what matters, of course; all ohmmeter ranges have 0 ohms at the right and infinity at the left. When you short the test leads together and the

meter reads 0 ohms, it's putting 125 milliamperes through R_2 , R_3 , and the test leads, but we know the meter has a 50-microampere movement, so nearly all the current must be going through the shunt (R_1).

Most likely, R_1 has decreased in value or there is a short circuit across it. Also, it looks as if R_2 has increased in value, or, more likely, you've set R_3 very high.

Look for solder bridges and defective switch contacts, then try replacing R_1 . You can find the correct value by trial and error; try 2 ohms as a first guess, and use a precision resistor for the final repair.

By the way, do not use the $R \times 1$ range to check circuits containing semiconductors or other delicate components—the heavy current can damage them. Use a digital meter for that.

Synchros and Selsyns And Accelerometers...Oh, My!

I just picked up some interesting 1970's military accelerometers. They consist of a fist-sized magnetically damped pendulum coupled to a synchro transmitter. These do seem to offer sub-g sensitivities. Yeah, solid-state accelerometers from Analog Devices (check out their ADXL05 in particular) or Motorola or others are getting fairly cheap, stable, and quite sensitive. However, these older mechanical beasts might still lead you to lots of interesting robot navigation, physics demos, racing car apps, or even military collectibles. With modifications, they might also make dandy inclinometers or levels. Figure 1 shows us what a typical unit looks like. More details on these are now posted to my www.tinaja.com/bargms01.html. All of which reminded me that it is way past time for us to look at...

Servos, Selsyns, and Friends

A lot of the stuff we routinely do with cheap electronics these days had to be handled at one time by special and expensive mechanical devices. A *servo* is any system that uses position or other feedback to gain accuracy. The feedback of most closed-loop systems improves precision but does so with price and stability problems. Usually, some position, velocity, or other factors is measured and

compared against your current goal—generating an error signal. The servo then seeks to minimize (or null out) errors. A servomechanism is a servo that involves mechanical motion of some

sort. A servomotor is just any motor or mechanical actuator that can be run in either direction and can be safely stalled in any position.

A *selsyn* is a system that consists of two or more motor-like devices that are intended to control or sense shaft positions. These might be true servo closed-loop or simple open-loop devices. Selsyns played a major role in all WWII ships and aircraft for position control. They are still used for ham antenna sensors, rugged industrial needs, or to measure wind speed and direction. Useful sources for selsyn parts are Fair Radio, Servo Systems, or C & H Sales.

Figure 2 shows a typical selsyn setup. A synchro transmitter consists of a three-phase rotary transformer. The single-phase rotor is driven from a 400-cycle (aircraft) or a 60-cycle (ship or land) source. Three sensing windings physically spaced 120 degrees around the stator output three sine waves whose values are related to the shaft angle. A second device

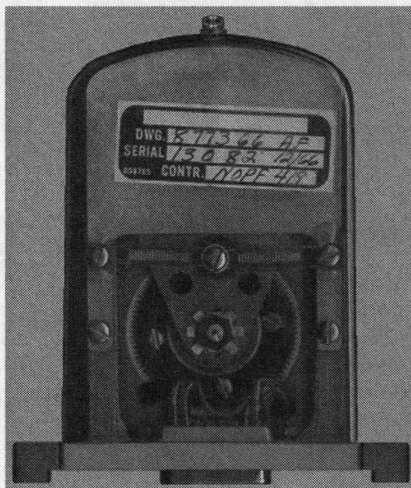


Fig. 1. An older military accelerometer that makes use of a pendulum coupled to a synchro transmitter is shown here.

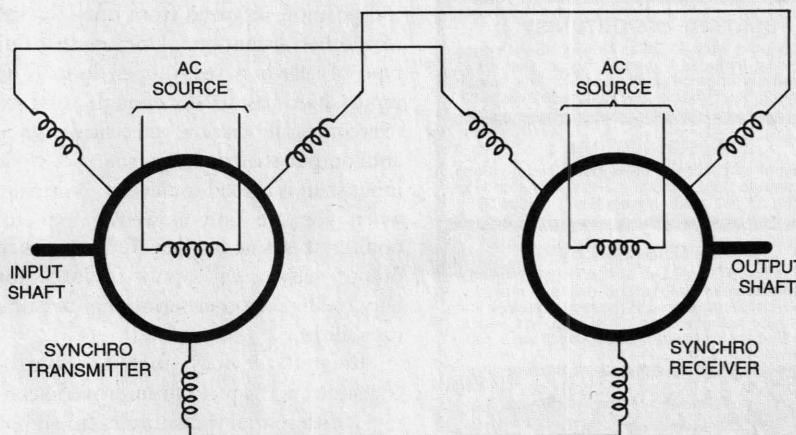


Fig. 2. A selsyn system consists of a synchro transmitter and a synchro receiver. The output shaft position follows the input.

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can be connected backwards across the three outputs. Its shaft will automatically follow the input shaft position. Thus, a pair of synchros could send any mechanical position or speed from one location to another distant one. Do note that this type of selsyn is really open loop. The input shaft has to do enough work to overcome all friction, electrical losses, and output-shaft-load inertia. So, your input shaft is loaded somewhat. Normally when you are only powering a meter pointer, this is no big deal. Other types of output selsyns will apply feedback for flap, rudder, or other serious tail-twisting capabilities.

Related devices include control transmitters (to pick up improved accuracy), differential transmitters (to add or subtract two shaft angles), or receivers (handy for shaft position displays). A *synchro resolver* does the calculations needed

to find an angle's X and Y components or to provide a digital display. Fancier selsyn combinations can do amazingly sophisticated analog trig calculations or similar tasks.

A useful synchro application-guide tutorial is available at www.litton-ps.com/DataSheetList.htm. Figure 3 is a photo of a standard synchro transmitter. Older sizes are usually measured in tenths of an inch. Thus a "size eight" selsyn is 0.8 inches in diameter while the "size fourteen" safely clears an inch-and-a-half space. Let me know if you need any of these to play with.

Dealing with 400 Hz

Military and other aircraft selected 400-Hz as an operating frequency because the size and weight (and thus the horsepower per pound) are much better than the usual 60-Hz utility

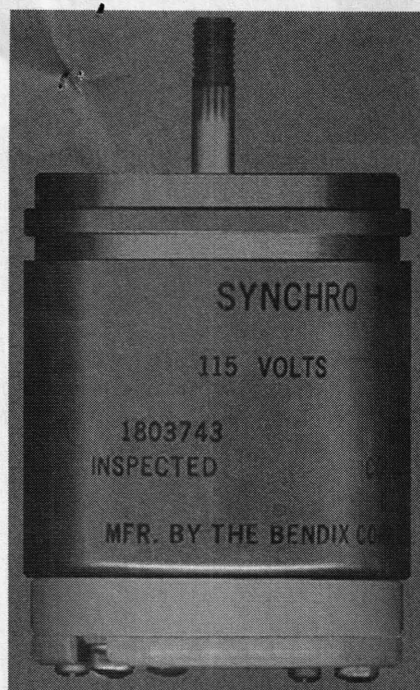


Fig. 3. A Synchro transmitter is used to sense shaft position.

power. The short transmission range of higher frequencies is not a factor in most airborne applications. Working 400-Hz power used to be a pain, but these days a PIC, some coils, and a few power semiconductors easily generate it. Mechanical "dynamotor" 60- to 400-Hz converters are readily available as military surplus at www.tinaja.com/barg01.html and a few other places.

This is a perfect way to apply my new magic-sinewave techniques that you can find described in detail at www.tinaja.com/magsn01.html. Can you run a 400-Hz device at 60 Hz? If you try this one-on-one, smoke and fire are certain to result. The lower winding inductance draws much higher current at a set voltage, saturates the iron, and burns up. Do not ever plug any 400-Hz device into the same voltage 60-Hz power line!

Few folks have noticed a sneaky workaround. Flux density is the name of the game. A 400-Hz device will usually run just fine on 60-Hz if you lower the voltage to 60/400ths or 0.15. The same current will produce the same magnetic flux, and the beast should remain happy. Naturally, the power and speed of a motor will be miniscule if you try this, but it is possible that it will still end up useful. A 400-Hz iron is often cheap.

Thankfully, in the case of a selsyn transmitter, these days you can use higher impedance electronic loads that you route to a PIC or whatever. A 12- or 15-

volt 60-Hz AC supply should work just fine. Thus, selsyn synchro transmitters may still be able to hold their own against encoders for some experimental applications—especially for measuring wind speed and position or for ham antenna sensing—and just may offer cost and reliability advantages.

Selsyns may seem dated, but servo and feedback concepts definitely are not. With this in mind, I have gathered together some of the better books on this subject for you as our resource sidebar. More details on all of these titles can be found at www.tinaja.com/amlink01.html.

Listening to Bats

Bats make use of utterly amazing ultrasonic-sonar systems. Their echolocation gets used both in a low-resolution mode for navigation and in high resolution to chow down on flying bugs. Considering the weight, power, resolution, and efficiency, a typical bat navigation system is some eleven orders of magnitude better than the best of military radar systems. Eons ago, the bats discovered the chirp secret to high-resolution radar. Namely, that a long swept frequency sine wave could be collapsed into the narrow pulses needed for high-resolution distance measurements. Collapsing can be done by taking the

Fourier transform of the swept return signal. However, bats more than likely use an acoustical-delay network having a linear time-versus-frequency delay characteristic. Details are shown in Fig. 4.

Intuitively, sending any swept FM sinewave into a linear time-delay versus frequency network stalls the higher frequency parts more than the lower ones, so they all pile up on top of each other and form a narrow pulse. Mathematically, you have done the frequency-domain to time-domain conversion through Fourier, Wavelet, or some similar transformation.

The sweep rates and frequencies depend on the species and whether the bat is in travel or chow mode. Typical sweep limits might be 20- to 40-, or 30- to 60-kHz, with repetition rates going as high as 200 per second. You cannot directly hear most of a bat's chirp, because of its ultrasonic frequencies and wide bandwidths. Instead, one of a number of sneaky schemes can be used to listen in. One of the simplest is pick up the bat's chirp with an ultrasonic alarm transducer. You then amplify and limit the signal and divide it down with a CMOS binary counter such as a 4040. The hard limiting removes many subtleties of the signal, but still lets you determine the presence and possibly the species of the bat.

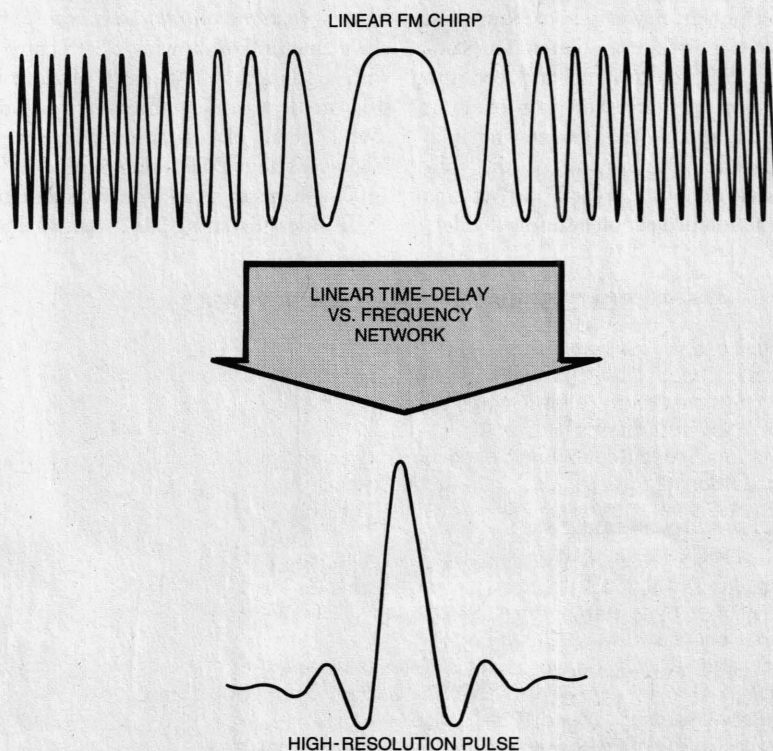


Fig. 4. A bat's chirp uses this sophisticated Fourier Transform scheme to dramatically improve its navigational resolution.


```
% POSTSCRIPT "404" ERROR EXTRACTOR/REPORTER
% =====
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% All commercial rights and all electronic media rights fully reserved.
% Personal use permitted provided header and entire file remains intact.
% Linking welcome. Reposting expressly forbidden.
% This PostScript-as-language routine reads an ISP log file and extracts all
% "404" or FILE NOT FOUND entries into a separate analysis file.
% To use this program, enter the full path sourcefile and target file names
% and keyphrase below and resave. Then distill the file.
% Note that a NO FILE PRODUCED message is normal and expected.
% IMPORTANT: Be sure to use "\\" when you mean "\" in any PostScript string!

/sourcefilename (C:\\medocs\\logfile.txt) def % sourcefile
/targetfilename (C:\\medocs\\le404file.txt) def % output file
/searchphrase (- 404) def % term to filter on
/workstring 2000 string def % string holds analysis line
/ws {writefile exch writestring} def % create output file object
/addtooutfile {ws (\\n) ws} def % add line to output file
/endoutfile {(\\n\\n) ws % close file when done
  writefile closefile} def

% checkline tests to see if a dash followed by a space and a 404 is present...
/checkline {dup searchphrase search {pop pop pop addtooutfile}{pop pop} ifelse} def

% /startoutfile creates an output file object...
/startoutfile {targetfilename (w+) file /writefile exch def(\\n\\nLines containing ) ws
  searchphrase ws ( in document ) ws sourcefilename ws (\\n\\n) ws} def

% Main loop reads one logfile line at a time for processing...
/grabphrase {sourcefilename (r) file /workfile exch def startoutfile
  {mark workfile workstring readline {checkline}{exit} ifelse
  cleartomark} loop endoutfile pop} def

grabphrase % This actually does it

%% EOF
```

Fig. 5. Postscript code to extract 404 log error messages.

The heterodyne methods can also be used if you down convert the bat's signal to the audio range where you can listen to it. A fancier, newer scheme known as *time dilation* works even better. You sample every 16th chirp signal and feed it to a dual-port memory. You then read out the signal at a 1/16 rate, converting it to audio while retaining all of the full-amplitude details. A great Tony Messina construction project on a simple divider-

style bat detector can be found at pw1.netcom.com/~t-rex/BatDetector.html.

Other technical bat resources are at life.csu.edu.au/batcall/abs/links.htm. The classic method of showing a bat's chirp is with a sonograph. An early pioneer in this field was Kay Elemetrics. Still found, of all places, at www.kayelemetrics.com. Their DSP Sonograph Model 5500-1 seems to be an industry standard.

Besides listening to bats, various

counting and monitoring schemes are often used. They can involve anything from clickety-clack manual counters to the most elaborate of 3-D multi-target image processing. Bat guano analysis can tell you a lot about pollution, climate, and other environmental changes as well. I am sometimes involved in an annual trek to Arizona's Falcon Creek Bat Cave to sample each year's guano deposits and return them for lab analysis. **WARNING: Before getting into this by yourself, be certain to read up on histoplasmosis.** Much more about Fourier and correlation appears in MUSE90.PDF and ACK54.PDF. More on CMOS counters and such can be found in my *CMOS Cookbook*.

The best place to go for accurate bat info is BCI, an acronym for Bat Conservation International. You can visit the Web site at www.batcon.org. For some really strange reasons, bat enthusiasts and cavers tend to go together. Maybe it is because they both sleep upside down. Thus, the National Speleological Society at www.caves.org also has all sorts of information and bat-related activities. Local chapters of the NSS are often called *grottos*. A typical state will have a dozen. Chances are that several are located near you.

Avoiding 404s

Preventing Web site errors can be a thankless and never-ending task. We saw last month how there is a great freebie Dr. HTML web service up at www.imageware.com/RxHTML or by way of my home page HTML button at www.tinaja.com. This is great for checking everything from links to spelling to programming syntax. We also saw how to extend this service to Acrobat .PDF files. A 404 or File Not Found error is any file that you cannot provide on your Web site. Some of these will be plain old mistakes visitors have made or else snooping attempts to hunt for pornography or to deduce your site structure. Others might be administrative related. For instance, the way my ISP has the site search set up, unnecessary errors are reported. The search succeeds and the visitor is happy, but the error is mysteriously generated anyway. A third source of errors is your own mistakes. These may be difficult or impractical to repair. They are often items such as an URL mention in some newsgroup, on a printed page, or a misquote. Nevertheless, the final error sources are your own mis-

SOME SERVO AND FEEDBACK BOOKS

Analytic Feedback System Design (Peter Dorato)
Automatic Control: The Power of Feedback (T. Djaferis)
Control System Design Guide (George Ellis)
Control System Dynamics (Robert Clark)
D.C. Motors Speed Control Servo Systems (R. Greene)
Design of Control Systems for DC Drives (A. Buxbaum)
Feedback Control of Dynamic Systems (Gene Franklin)
Feedback Control Systems (Charles Phillips)
Force and Touch Feedback for Virtual Reality (G. Burdea)
Introduction To Feedback Control Systems (Pericles Emanuel)
Introduction to Servomechanism System Design (W. Humphrey)
Modern Control Systems (Richard Dorf)
Nonlinear and Optimal Control Systems (Thomas Vincent)
Principles of Process and Servo Control (D. Senko)
Quantitative Feedback Theory (Constantine H. Houppis)
Schaum's Outline of Feedback and Control Systems (Schaum's)
Synchro and Resolver Conversion (Geoffrey Boyes)
 For more book details, see www.tinaja.com/amlink01.html.

takes that you can correct.

These you must ruthlessly stomp out when and as you find out about them. You find your errors by reading the raw log files that your ISP should be providing you. Note that most statistics software only does error summaries instead of providing the needed details. Errors can be found by searching the log files for a "404." Of special interest is their referral information in the same line. This tells you where your visitor came from. When they arrived from another of your pages, you have a great clue to exactly when and where the error came down. A reasonable goal should be to keep your 404-error rate under one half of one percent.

Our PostScript-as-language code for this month reads your log file, extracts the useful parts of each 404-error line and reports them back to you. This can be a convenient way to improve your web stats. Details are shown in Fig. 5.

You should first enter a PostScript program into a word processor or text editor and modify it for your intended use. You then save the file as a standard ASCII text file. The file is then sent to Acrobat Distiller or GhostScript. Extracted output error lists are reported onscreen in the log file and are saved as separate files for later analysis and correction. More on PostScript-as-language in www.tinaja.com/post01.html.

New Tech Lit

Microsoft has recently upgraded their already outstanding TerraServer site. Topographic maps have been added to their aerial photos. They are also in the process of adding bunches of photo coverage with accuracy to one meter. You can find all this up at www.terra server.microsoft.com. On the other hand, you can just click on the AERIAL button up at my www.tinaja.com. A monster "my-database-is-bigger-than-yours" feud seems to be going on between the TerraServer folks and the online IBM patent resource up at patent.womplex.ibm.com. Both of these clearly seem to be moving up well into terabyte territory. More on the patently absurd is up at www.tinaja.com/patnt01.html.

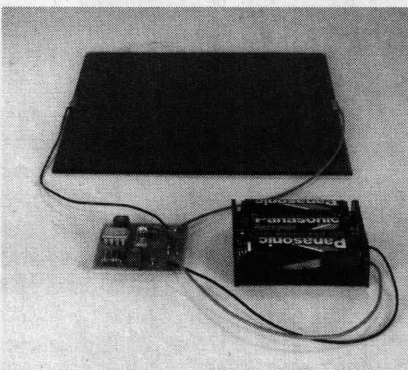
From Micro Linear comes a new CD on their integrated circuit line. Useful application notes are included on lamp ballast and intelligent battery management. Free info on CAD (computer-aided design or drafting) can be found up at www.cadprimer.com. Engineering links

are provided for you at www.spacey.net/davis/frames.html. Or hit the LEROY button on my home page at www.tinaja.com. The SETI-at-home project has now shattered all records for the largest computing project of all time. It is now racking up over one millennia per day of computing time! You are invited to become a participant by picking up details at setiathome.ssl.berkeley.edu.

Our featured trade journals for this month: *BookTech*, *Print on Demand Business*, and *Digital Graphics*, with coverage of emerging and exciting new self-publishing opportunities. More information can be found at www.tinaja.com/bod01.html. **P**

NOCTURNAL FLAME

(continued from page 28)



The completed Nocturnal Flame is ready to capture the sun's energy for you.

type of cell, solder two-lengths of wire to the outside of each channel clip. Scrape the silver backing from the back of PC1 where the channels will attach. Bend the edges of the channels slightly inward so that they will fit snugly on the solar cell and slide them on. Remember, solar cells are fragile; be careful not to break the glass.

When the Nocturnal Flame is completely assembled, check your work over for errors such as bad solder joints, wrong components, or other similar mistakes that seem ridiculous or embarrassing once they are found.

Place PC1 in a window or under a bright light. Turn R5 and R6 fully counterclockwise. That will set the PWM level to 0 and disable LED1. Measure the voltage from the solar cell. If the light is bright enough, you should at least see a couple of volts. Also measure the battery volt-

age at the positive terminal of B1. You should see 0.7 volts less than the solar cell's output voltage. Depending on the state of the batteries and the intensity of the light source, there should be a couple of volts, and it should be rising by a few millivolts very slowly. Keep the cell under the light for 10 minutes or so until the batteries achieve a partial charge. Now turn R6 fully clockwise. That will set the PWM level of LED1 to full. Turn R5 slowly clockwise. If everything went well, LED1 should come on somewhere before the pot hits the clockwise stop. If it did not come on, check the component placements and connections.

Operation. The operation of the Nocturnal Flame is very straightforward. Point PC1 at the sun to charge B1 and wait until dusk. When it is dark enough for you to want LED1 to come on, adjust R5 clockwise until LED1 lights. Adjust R6 for the desired brightness level.

Build one or more Nocturnal Flames and use them as accent lights around your deck, to highlight steps or paths around your home, as night lights in your home, or even just for custom lighting effects. **P**



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MA02

BUILD A SUPER-SENSITIVE HEAT DETECTOR



SKIP CAMPISI

Measuring and sensing temperature has resulted in many different methods and devices. While some of the solutions developed meet certain requirements such as temperature range and accuracy, most have some sort of characteristic that makes them unsuitable for the average home hobbyist.

Sensors such as thermocouples, thermistors, and exotic ICs usually require their own "flavor" of circuitry in order to use their outputs. The outputs are also usually non-linear and change only slightly in relation to the temperature change. The result is a system that can easily be inaccurate because of circuit noise being interpreted as temperature change.

Fortunately for the home experimenter, materials and components are readily available for building an inexpensive non-contact type of temperature sensor that overcomes the limitations listed above. Such a device is presented here. The unit actually indicates the temperature *difference* between the temperature of an object and an arbitrary "ambient" temperature reference that can be set at any time.

A gain control is used to adjust the sensing range from 1°C down to an incredible 0.01°C between the sensor and the ambient reference temperature! While that specification might sound like "design overkill," it is really quite essential for remote sensing. For example, a match flame can be detected up to a distance of one foot at the 1°C gain setting. With the gain increased

*Detect hot or cold objects
remotely with this
ultra-sensitive
infrared-sensing device*

to detect a difference of 0.1°C , the heat from the palm of your hand can be detected at the same distance. At the maximum gain setting, that same match flame can be detected from across a room!

Obviously, those objects are all more than 1°C hotter than room temperature, so why is such a high gain needed? The culprit is, of course, the "column" of air between the object and the sensor. As the air absorbs and dissipates some of the heat, the apparent temperature falls off rapidly in an inverse-square relationship with the distance—double the distance, and the heat energy drops by a square-root function. By the time the heat reaches the sensor, we are probably sensing convection currents more than the actual, direct heat from the object.

The implication is that any air flow will disturb the readings. When using the device at higher gains, still air is the biggest requirement for consistent results. Avoid fans, heat vents, and the like when experimenting. Of course, that "disadvantage" can

also become an advantage, especially when using the unit to *detect* air flow such as "leak" testing.

About the Circuit. As you can see on the schematic diagram in Fig. 1, the actual circuit for the Heat Detector is a deceptively simple design. Remember, however, that the signals involved with sensing a heat source at a distance are very low in amplitude. Low-amplitude signals always run the risk of being lost in the normal background noise that they are sitting in, so all of the components were selected for low-noise performance. For example, notice that all of the fixed resistors are 1% metal-film units. Those types of resistors perform extremely well when low noise is important. What is interesting in the overall design is that the actual values of the resistors are *not* critical—any value within 5% or even 10% of those given should work well. However, the metal-film composition of the resistor is mandatory!

The circuit is a "current mirror" stretched to the ultimate degree. The current mirror itself is built around Q1 and Q2, a pair of surface-mount transistors. The tiny mass of the SOT-23 transistor package lets the circuit respond extremely fast to temperature changes.

A standard current-mirror circuit would only have Q1 and Q2 with their bases and emitters connected in parallel and the base of Q1 shorted to its collector, forming a diode-connected transistor. The current

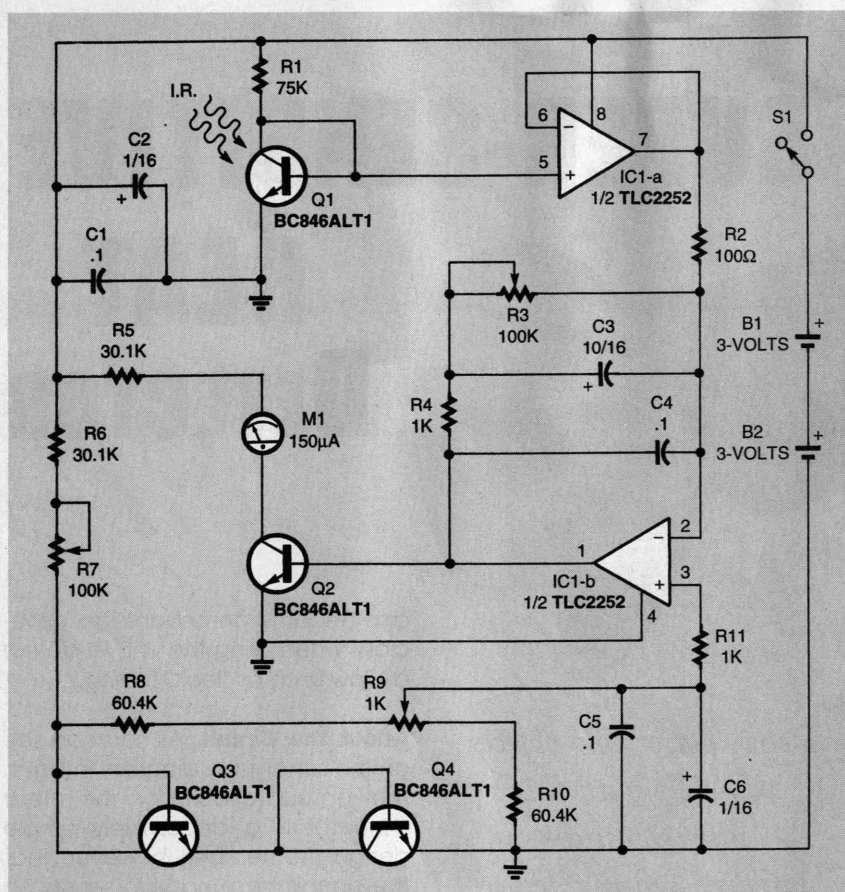


Fig. 1. The Heat Detector is built around a simple current-mirror circuit. When one transistor is heated up, the change in its current flow is reflected in a parallel circuit.

flowing through R1 would then be "reflected" in Q2's collector. For both currents to be equal in size and temperature tracking, Q1 and Q2 are normally a matched pair in a single package. However, the characteristic that we are looking for is the temperature difference between Q1 and Q2; that is why individual units are used. For every 1° C increase in the junction temperature of a silicon transistor at a fixed bias current, its forward-bias voltage decreases by about 2 millivolts. By holding Q2 at a fixed ambient temperature, its collector current will decrease as Q1's temperature rises.

A change of only 2 millivolts is not a lot to work with in a remote-sensing circuit, especially if we are interested in displaying a difference of 0.01° C. Besides, an increase in current for an increase in temperature would be very desirable.

Placing Q1 outside of the cabinet at the focal point of a standard parabolic flashlight reflector makes the sensor very directional and

extremely sensitive to infrared energy focused by the reflector. The rest of the components are located inside a fully-enclosed cabinet that is used as an ambient-temperature reference.

Between Q1 and Q2, a two-stage inverting variable-gain amplifier made up of IC1-a and IC1-b provides the high gain needed for remote sensing. The first stage (IC1-a) is a unity-gain buffer that drives the second stage (IC1-b). The second stage is an inverting amplifier with a gain that can be adjusted between about -10 and -1000 with R3. The audio taper of R3 makes adjusting easy; the most useful gain in setting, -100, is at about the center of its rotation.

A 1.2-volt reference, built around Q3 and Q4, tracks temperature changes within the case at a rate of -4 millivolts per degree C. The reference is divided exactly in half, using R8-R10, to a 0.6-volt reference tracking at -2 millivolts per degree C. That output is applied to

the reference input of IC1-b. Note that the bias resistors for the transistors have been chosen so that Q1-Q4 all operate at the same current level. That will help ensure that their forward voltages are approximately equal, assuming that they are at the same temperature.

The 0.6-volt reference acts as a "ground" reference for the input signal from Q1. When Q1 heats up by 1° C, its forward voltage decreases by 2 millivolts. With the reference set by R3 to its minimum gain of -10, the voltage appearing at Q2's base and emitter increases by 20 millivolts, which about doubles its collector current. The result is a full-scale display on M1. At a gain of -1000, a 0.01° C increase results in a 20 microvolt signal giving a full-scale display! With Q1 at ambient, Q2 conducts enough current for a 50%-of-full-scale display on M1.

As you might suspect, setting the operating levels of the transistors is crucial in order to get any meaningful results out of the instrument. That is done by leaving Q1's current fixed and varying the current to Q3 and Q4 with R7. When adjusting the 1.2 volt reference, R7 can be considered as a "coarse" adjustment. Any "fine" adjustments are then done with R9.

The collector of Q2 directly drives M1, an analog current meter. Although the circuit uses a 150-microamp unit, you can use any analog meter that has a full-scale display between 50 and 250 microamps. Keep in mind that the markings on the meter do not indicate any particular temperature reading, just the relative change. If you are going to use a meter of a different capacity, R1, R5, and R6 must be changed in order for the circuit to work. The different values for those components are shown in Table 1 for several different types of meters. Those resistors have been selected to provide current through the transistors so that they all operate at about 50% of M1's full-scale current. The meter is protected from excess current by R5.

The total current drain of the circuit is only about 300 microamperes when M1 is a 150-µA unit. A simple unregulated power supply is quite suitable in that case. Two 3-volt lithi-

um coin cells are connected in series for a 6-volt supply. With a rating of about 200 milliamp-hours, they will last a long time. They also have a very "flat" discharge voltage curve, simulating a regulated power supply at those low currents.

One final consideration for this type of circuit is the concern over external noise pickup, such as 60-Hz hum and radio-frequency interference (RFI). Capacitors C1-C6 were specially selected to filter out as much noise as possible while still keeping response times very fast. Tantalum capacitors are highly recommended. If you have noise problems, the values of the capacitors can be increased. You can even connect the parabolic reflector to ground through one of its mounting screws if necessary.

Construction Tips. The Heat Detector's circuit is simple enough that it can be built on perfboard using standard construction techniques. One area that will be a bit more difficult is connecting the tiny surface-mount transistors to the circuit. The simplest method is to attach separate leads directly to the units, converting them to through-hole-like devices. They can then be mounted to a perfboard like any other standard leaded part. The technique is shown in Fig. 2. Use 30-gauge wires that are at least 1-inch long for the leads—the excess length can be trimmed off after the unit is mounted to the board. In order to help protect the transistor from heat-related damage during soldering as well as making it easier to handle the tiny parts, hold the transistor case with a "micro-gator" clip such as a RadioShack 270-373. That particular clip has flat jaws rather than serrated teeth.

Using small tweezers, carefully bend the transistor leads out flat from the case. Wrap a turn or two of wire around the unit's leads with the tweezers. Put the clip in a bench vise and attach another clip to the free end of the wire so that it acts as a weight, stretching the wire. Using very fine solder and a very fine tip on your iron, carefully solder the wire to the lead. Use as little heat as possible for a very brief time in order to avoid damaging the transistor.

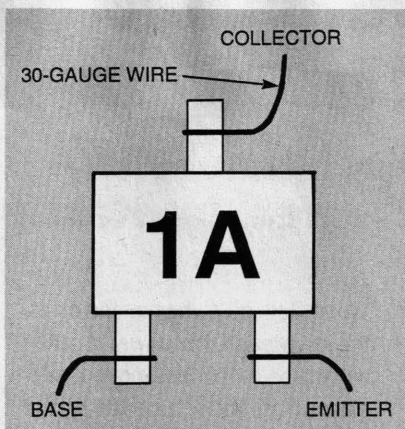


Fig. 2. Surface-mount transistors can be converted to use with perfboard with some careful work.

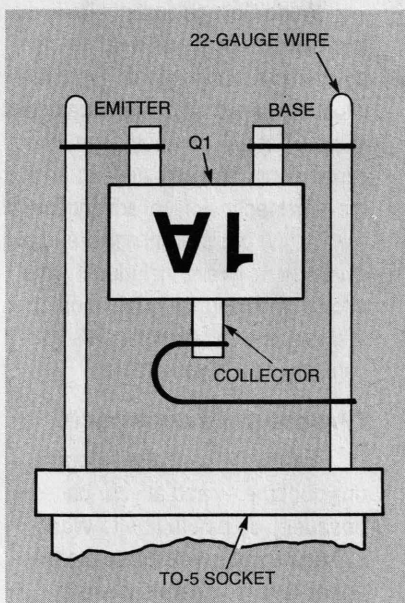


Fig. 3. Transistor Q1 is mounted to a TO-5 socket so that it can be moved to the focal point of a parabolic reflector.

Repeat the procedure for all three leads on all four transistors. Once the solder joints are cool, check the transistors with an ohmmeter to verify that the transistors were not damaged during soldering.

The BC846ALT1 transistors specified for Q1-Q4 are general-purpose NPN small-signal units. Any similar type of device can also be used. Examples of alternate part numbers include BC848ALT1, MMBT3904LT1, and MMBT4401LT1. Note, however, that it is important that all of the transistors are the same device type in order for the correct matching of transistor characteristics. In other words, don't mix and match when selecting parts for Q1-Q4.

When laying out the components

on the perfboard, it is a good idea to place R3, R7, and R9 along one side so that their controls point in the same direction. Since R7 and R9 are 20-turn units having only a small adjustment screw, it would be convenient to use some sort of panel-knob adapter for them. Such a device is available from Spectrol for their line of multi-turn potentiometers. Other manufacturers might have similar devices available. The adapters should be mounted to the potentiometers according to the manufacturer's instructions.

Transistors Q3 and Q4 should be installed between R7 and R9 with their cases sitting about 1/4-inch above the board for good air circulation around the transistors. Mount Q2 in a similar fashion. Remember that Q1 will not be mounted on the board; it will be discussed later.

Next, select a plastic cabinet that is large enough to hold M1, S1, and the completed circuit board. Cut an appropriate hole for M1; locate and drill the three mounting holes for R3, R7, R9, and S1. The board will be supported by the three potentiometers.

A parabolic reflector from a flashlight or lantern will be used for the collector. Although the size can vary depending on what you have available, a 3-inch-diameter reflector from a 6-volt lantern is a very effective size for the intended use.

Select a transistor socket that will fit through the light-bulb hole at the base of the reflector. A standard TO-5 or TO-39 socket should be fine in most cases. However, it might be necessary to use a smaller TO-18 socket. Install the socket in a hole centered in one end of the cabinet, using silicone sealant to hold it in place if necessary.

Drill a couple of clearance holes through the base of the parabolic reflector on either side of the light-bulb hole for a pair of 2-56 or 4-40 mounting screws. Place the reflector over the transistor socket and mark and drill the screw holes.

Using Fig. 3 as a guide, insert two 1-inch lengths of 22-gauge tinned solid-bus wire into the two widest-spaced holes in the transistor socket. Install Q1 by carefully wrapping its wire leads from its base and

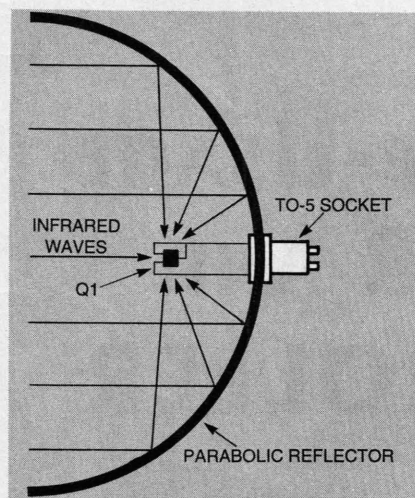


Fig. 4. When Q1 is at the focal point of a parabolic reflector, all of the gathered heat energy will strike the transistor. The size of the reflector will increase the ability of the Heat Detector to collect energy from the source of heat, increasing the gain of the system.

PARTS LIST FOR THE HEAT DETECTOR

SEMICONDUCTORS

IC1—TLC2252 CMOS dual op-amp, integrated circuit (Texas Instruments)
Q1-Q4—BC846ALT1 Surface-mount general-purpose transistor, NPN (see text)

RESISTORS

(All resistors are 1/4-watt, 1% metal-film units, unless otherwise noted.)
R1—75,000-ohm (see text)
R2—100-ohm
R3—100,000-ohm potentiometer, audio-taper
R4, R11—1000-ohm
R5, R6—30,100-ohm (see text)
R7—100,000-ohm potentiometer, 20-turn (Spectrol)
R8, R10—60,400-ohm
R9—1000-ohm potentiometer, 20-turn (Spectrol)

CAPACITORS

C1, C4, C5—0.1- μ F, monolithic ceramic
C2, C6—1- μ F, 16-WVDC, electrolytic
C3—10- μ F, 16-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

B1, B2—3-volt lithium battery (CR2032 or similar)
M1—Analog panel meter, 150- μ A (see text)
S1—Single-pole, single-throw switch
3-inch parabolic flashlight reflector, enclosure, battery holder, wooden or plastic handle, knobs, shaft adapters for R7 and R9, hardware, etc.

emitter around the 22-gauge wires as shown—do not solder the connections just yet.

Carefully slide Q1 on its leads to about the center of the 22-gauge bus wires. Attach the reflector to the case without disturbing Q1. With everything assembled, Q1 should be at a point close to the focal point of the reflector, similar to the arrangement shown in Fig. 4. It does not have to be exact at this point; we will next be adjusting the assembly's focus. Set the reflector assembly down on a table or bench so that you can look directly down the "business" end of the reflector at Q1.

Starting at a distance of about a foot or two away from the reflector, look at the "image" of Q1 in the reflector. You should see a black "ring" somewhere around the reflector's inside circumference, either towards the front or rear. The black ring is the focused image of Q1's case. If you don't see the ring, try sliding Q1 further in or out and look again; it might take a little practice. What you see should look similar to the view shown in Fig. 5.

Once you see the black ring, move farther away from the assembly while keeping the ring in sight. What you want to end up with is the black ring just about filling the forward (widest) section of the reflector from a distance of around 8 to 10 feet. That is a good focal point for indoor use of the Heat Detector. Slide Q1 in and out a little at a time to achieve that focus.

You can confirm a good focus by pointing the reflector directly at a diffuse light source across the room such as a lamp with a shade or a sun-lit window curtain. Peek over the edge of the reflector, and you should see Q1 glowing like a lamp as the focused light shines on it. DO NOT point the unit directly at the Sun or you might melt or severely damage the transistor!

Once you have a good focus, carefully remove the reflector and solder the base and emitter wire leads of Q1 to the 22-gauge wires. Wrap the collector wire lead of Q1 around the 22-gauge wire connected to Q1's base lead and solder that connection also. Re-check the

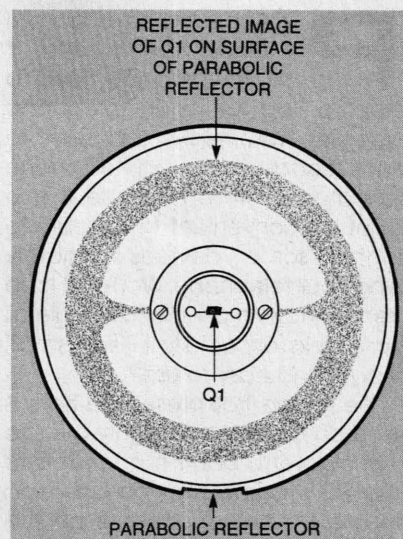


Fig. 5. When focusing Q1, you should see a black band around the surface of the reflector. Careful adjustment of Q1 can vary the apparent size of the reflector to the rest of the circuit.

TABLE 1

ALTERNATE RESISTORS FOR M1

M1	R1	R5	R6
50-mA	215k	90.9k	100k
100-mA	110k	45.3k	49.9k
150-mA	75k	30.1k	30.1k
200-mA	53.6k	22.1k	20k
250-mA	43.2k	18.2k	15k

focal point; if everything looks good, cut off the excess 22-gauge wires.

Select a non-metallic handle, such as a plastic or wooden one, and mount it to the opposite end of the case. It is important that the handle not conduct any heat from your hand, or the accuracy of the unit will be affected. If you use a handle from a wood file, the threads in the handle will make it easier to mount the handle to the case.

Now you can finish up assembling the Heat Detector. Install the board, S1, and M1, and then wire up the rest of the circuit. If you want to label any of the controls, that should be done before installing any other parts. When wiring Q1 into the circuit, pay careful attention to the polarities of that component.

Install the reflector, the handle, and close up the case. Install the knobs on the potentiometers and set the unit aside for at least 15 minutes to let the temperature inside stabilize after all of the handling it just received. You are now ready to test.

(Continued on page 50)

to know how to program in QBasic, but it's very easy to modify the code and see what happens. Many budding programmers start gaining hands-on experience by modifying a program to achieve a particular goal. Just make sure that you have a backup copy of the program before you start experimenting. The program is well documented and will stand alone to act as a guide for those who are familiar with QBasic.

One example of a need to modify the program has to do with distinctive-ring service. The program will detect and count a normal ring cycle (two seconds of ringing followed by four seconds of silence). Distinctive ringing might cause the Phone Troll to count each burst of rings, indicating two or three rings for each actual ring. Simply divide the variable *ringct* by two or three, or rewrite the program to indicate which sequence of rings was received.

Like any good servant, your troll awaits. Ω

HEAT DETECTOR

(continued from page 41)

Using the Heat Detector. There are a few points to consider whenever using the Heat Detector to obtain consistent readings: always grasp the unit by the handle only, keeping your body heat as far away from the cabinet as possible. For critical operation, attaching the unit to a camera tripod would be an excellent method.

Unless you are actually leak-testing and looking for circulating air, any air that blows across Q1 will cause temperature drift in your readings. Still air is the priority here.

The parabolic reflector used is very directional and will take some practice in aiming. Pointing a standard flashlight at an opposite wall will give you an idea of the area of coverage to be expected.

Hold the unit by the handle and set R3 to its *minimum* gain setting. Set R7 and R9 to their mid-positions. Turn on the power and point the reflector at a blank wall.

Adjust R7 for a display of about half of full scale on M1. That is the most sensitive area, although any

region between 20% and 80% will work well. If R7 won't get you on scale, it might be necessary to adjust R6's value. An increase in value lowers M1's reading, and vice versa.

Advance R3 slowly while watching the display—it will probably move off-scale in one direction or the other as you arrive at the maximum gain setting. That is due to inherent offset voltages in the components. Adjusting R7 should give full-scale coverage in either direction at the highest gain setting. If not, adjust R6's value as mentioned above.

Note that making any settings with R7 becomes more difficult as the gain is increased. That is where R9 becomes very useful. For best results, always adjust R7 first to get as close as possible; then use R9 as a "fine" adjust for a 50% of full-scale indication on M1.

Set R3 back down to its minimum gain setting and adjust R7 for a 50% reading on M1. Now that you have an ambient-temperature set-point from the blank wall, point the unit at a match or candle flame about 6 to 12 inches away. You should immediately "peg" M1. Return to the blank wall and the display will return to 50%.

Rotate R3 to mid-position, and adjust R7 for a 50% reading on the blank wall. Hold the palm of your hand about 6 to 12 inches in front of the reflector and M1 should go full-scale again. Hold an ice cube in front of the reflector and M1 should drop to its minimum reading.

Rotate R3 to its maximum position, and adjust R7 (followed by R9) for a 50% reading on the blank wall. Now you can "sweep" slowly around the room. A hot spot such as a sun-lit window or a lamp will probably peg M1 at full-scale. You will now note that any breeze, even one caused by rapid movement of the Heat Detector, will cause display drift!

Going Further. You will by now have realized that the parabolic reflector plays a critical role in the sensitivity and area of coverage of the Heat Detector. A 3-inch-diameter reflector is an excellent size for general-purpose use. However, it can be difficult to aim accurately at distant

objects.

The solution to that problem is to use a larger reflector. Any photographic supply store will carry larger-diameter reflectors. Of course, the sensitivity of the instrument increases with the square of the diameter in addition to the coverage.

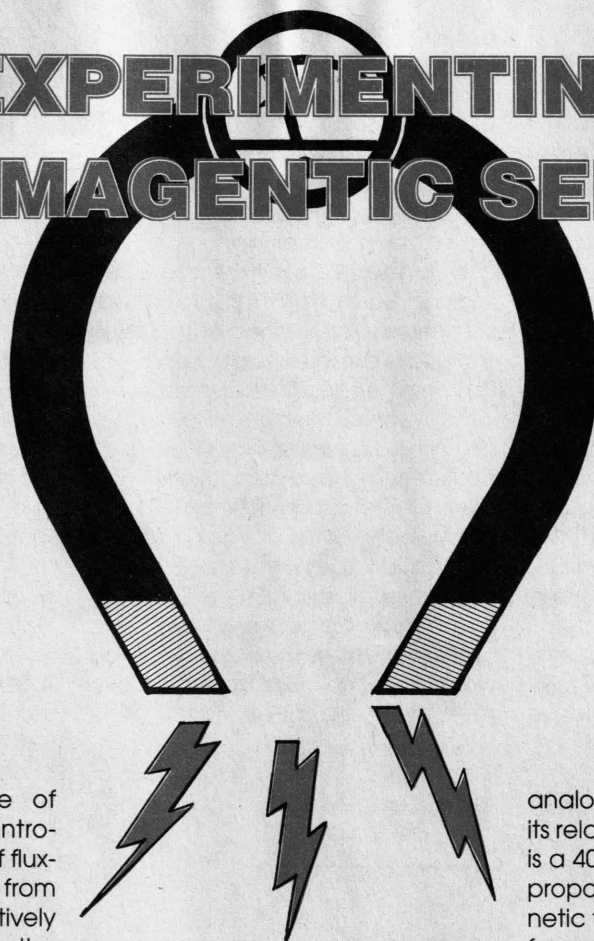
Some applications such as remote testing of circuit boards for hot spots might have the opposite requirement—miniature parabolic reflectors. A 1-inch-diameter reflector would be highly directional. The Heat Detector has sufficient gain to compensate for the decreased gain from the reflector. A miniature flashlight would be a good source for smaller reflectors.

Of course, you still have to make sure that Q1 is focused properly. An adjustable focus arrangement would be a great help in close-up work, where the focal point moves further away from the rear of the reflector. One method that can be used to make the Heat Detector more versatile would be to use a TO-5 transistor socket with female pins that are spring-loaded and completely hollow through their length. That would allow you to slide Q1 back and forth to achieve any focus position.

The Heat Detector is an easy-to-build precision instrument that shouldn't cost more than about \$25 to put together even using all new parts. Any salvaged or surplus components will drop the cost of the project to the point where it becomes an almost trivial matter to add such an instrument to your bench or tool kit. Build one yourself and have hours of educational fun, or even put it to some practical use! Ω



EXPERIMENTING WITH MAGNETIC SENSORS



In the June, 1998 issue of **Electronics now**, we introduced the FGM-x series of flux-gate magnetometer sensors from Speake & Co. Ltd. These relatively low-cost devices are made in the United Kingdom, but are distributed in the United States by Fat Quarters Software (24774 Shoshonee Drive, Murrieta, CA 92562; Tel: 909-698-7950; Fax: 909-698-7913; Web: www.dconn.com/fatquarterssoftware). For those not with us then, let's quickly review some of the important features of the FGM-3, which is used in the projects that are described in this article.

The FGM-3 is a three-terminal device (see Fig. 1A) that will measure magnetic fields over the range ± 0.5 oersted (± 50 mteslas). It can be used in a wide variety of magnetometer and gradiometer circuits. The three terminals on the FGM-3 are: RED: POWER (+5 VDC); BLK: GROUND (0V); and WHT: SIGNAL.

The output signal is a train of TTL-compatible pulses with a period that ranges from 8 to 25 mS, or a frequency range of 40 to 125 kHz. The magnetic field strength is indicated by the frequency produced, and can be read on a variety of display devices including digital frequency counters, digital period counters, analog meters, and even

*This month, we put our
flux-gate sensors to work
and build a practical
magnetometer and
gradiometer.*

JOSEPH J. CARR

computers. It is relatively easy to interface the FGM-3 device to microcontrollers—such as the Parallax BASIC-Stamp or Micromint PicStic devices—using a line of interface chips offered by Speake and Fat Quarters.

The sensitivity pattern of the FGM-3 device is shown in Fig. 1B. It is a figure-8 pattern with its maxima along the major axis of the FGM-3, and minima orthogonal to the major axis. At other angles, the sensitivity drops off as the cosine of the angle from the major axis.

Analog Interface to FGM-3. Figure 2 shows a method for providing an

analog interface to the FGM-3 and its relatives. The output of the sensor is a 40- to 125-kHz frequency that is proportional to the applied magnetic field. As a result, we can use a frequency-to-voltage (F/V) converter such as the LM2917 to render the signal into a proportional DC voltage. That voltage, in turn, can be used to drive an analog or digital voltmeter or milliammeter. The LM2917 is used here because it is widely available at low cost from mail-order parts distributors.

The output circuit consists of a bridge made up of R1-R3 and the output of the LM2917. A resistive voltage divider (R2/R3) produces a potential of $\frac{1}{2}(V_+)$ at one end of the 22,000-ohm sensitivity control (R4). If the voltage produced by the LM2917 is the same as the voltage at R4, then the differential voltage is zero and no current flows. But if the LM2917 voltage is not equal, then a difference exists and current flows in R4 and M1. That current is proportional to the applied magnetic field. Meter M1 and R4 can be replaced with a digital voltmeter, if desired.

The DC power supply uses two regulators, one for the FGM-3 and one for the LM2917. Even better results can be obtained if an intermediate voltage regulator, say a 78L09, is placed between the V_+

source and the inputs of IC2 and IC3. That results in double-regulation, and produces better operation.

Digital "Heterodyning." A different interface method is shown in Fig. 3. It results in a more sensitive measurement over a small range of the sensor's total capability. In other words, that circuit makes it possible to measure small fluctuations in a relatively large magnetic field.

In that circuit, a D-type flip-flop is used to "mix" the frequency from the FGM-3 with a reference frequency (F_{REF}). The FGM-3 literature calls that process "digital heterodyning," although it is quick to point out that it is really more like the production of alias frequencies by under-sampling than true heterodyning.

Two types of frequency sources can be used for F_{REF} . For relatively crude measurements, such as a

passing-vehicle detector, the CMOS oscillator of Fig. 4A is suitable. This circuit is based on the 4049 hex inverter connected in an astable multivibrator configuration. The exact frequency can be adjusted using R2, a 10,000-ohm, 10-turn potentiometer.

Where a higher degree of stability is needed, for example when making Earth-field variation measurements or testing magnetic materials, a more stable frequency source is needed. In that case, use a circuit such as the one in Fig. 4B. That circuit uses a crystal-controlled oscillator feeding a binary divider network. Crystal oscillators can be built, or if you check the catalogs you will find that a large number of frequencies are available in TTL- and CMOS-compatible formats at low enough costs to make you wonder why you would want to build your own. I've seen them sold for about the price

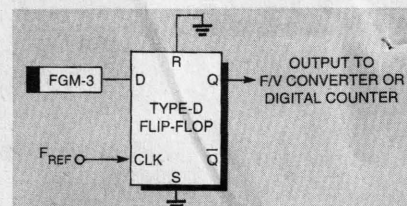


Fig. 3. To provide a more sensitive measuring capability, but over a smaller range, a type of "digital heterodyning" can be used.

of a crystal alone in many mail-order catalogs (Digi-Key, etc.).

The reference frequency is adjusted to a point about 500 Hz below the mean sensor frequency. That frequency is measured when the sensor is in the east-west direction. That arrangement will produce a frequency of 0 to 1,000 Hz over a magnetic field range of ± 500 gamma.

A Magnetometer Project. Figure 5 shows the circuit for a simple magnetometer based on the FGM-3 flux-gate sensor. A PC board, the sensor, the ICs, and most other parts (but no power supply, enclosure, or switch S1) can be obtained from Fat Quarters Software (contact them directly or see their Web page for more information). That circuit takes the output frequency of the FGM-3, passes it through a special interface chip (IC1), and then to a digital-to-analog converter to produce a voltage output.

The heart of the circuit, other than the FGM-3 device, is the special interface chip, Speake's SCL006 device. It provides the circuitry needed to perform magnetometry, including Earth-field magnetometry. It integrates field fluctuations in one-second intervals, producing very sensitive output variations in response to small field variations. It is of keen interest to people doing radio-propagation studies, and those who need to monitor for solar flares. It also works as a laboratory magnetometer for various purposes. The SCL006A is housed in an 18-pin DIP IC package.

The D/A converter (IC2) is an Analog Devices type AD557. It replaces an older Ferranti device seen in the Speake literature because that older device is no longer available. Indeed, being a European device, it was a bit hard to find in unit quantities required by

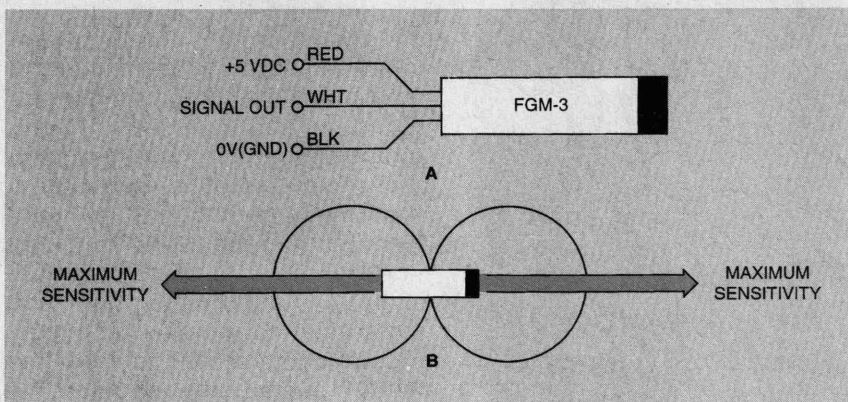


Fig. 1. The FGM-3 sensor is a three-terminal device that can be used to measure magnetic fields (A). Its sensitivity pattern is a figure-8(B).

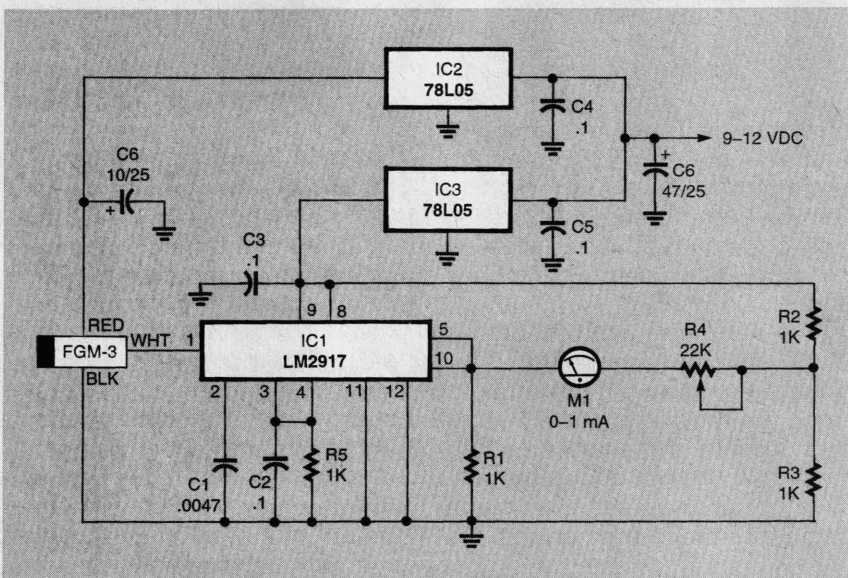


Fig. 2. This circuit can be used to provide an analog interface to the FGM-3. The heart of the circuit is an LM2917 F/V converter, which is inexpensive and easy to find.

hobbyists on this side of the Atlantic. The kit from Fat Quarters Software contains all the components needed, plus a printed circuit board. The FGM-3 device is bought separately.

The circuit is designed so that it could be run from 9-volt batteries for use in the field. A sensitivity switch, S1, provides four positions, each with a different overall sensitivity range. The output signal is a DC voltage that can be monitored by a strip-

chart or X-Y paper recorder, voltmeter, or fed into a computer using an A/D converter.

If you intend to use a computer to receive the data, then it might be worthwhile to eliminate the D/A converter and feed the digital lines (D0-D7) from the SCL006A directly to an eight-bit parallel port. Not all computers have that type of port, but there are plug-in boards available for PCs, as well as at least one

product that makes an eight-bit I/O port out of the parallel printer port.

Gradiometers. One of the problems with magnetometers is that small fluctuations occur in otherwise very large magnetic fields, and those fluctuations can sometimes be important. A further problem with single-sensor systems is that they are very sensitive to orientation. Even a small amount of rotation can cause unacceptably large, but spurious, output changes. The changes are real, but are not the fluctuations that you are seeking.

A gradiometer is a magnetic instrument that uses two identical sensors that are aligned with each other so as to produce a zero output in the presence of a uniform magnetic field. If one of the sensors comes into contact with some sort of small magnetic anomaly, then it will upset the balance between sensors, producing an output. The gradiometer gets its name from the fact that it measures the gradient of the magnetic field over a small distance (typically 1 to 5 feet).

These instruments can be used for finding very small magnetic anomalies. For example, the metallic firing pin of plastic land mines buried a few inches below the sur-

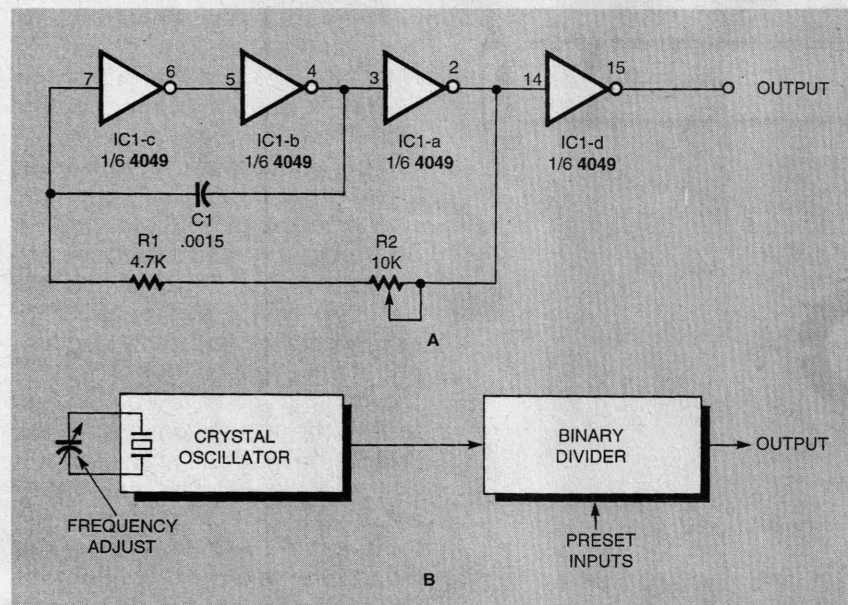


Fig. 4. For relatively crude measurements, a CMOS oscillator like the one in A can be used to generate the reference frequency. In more demanding applications, a crystal oscillator and binary divider can be used (B).

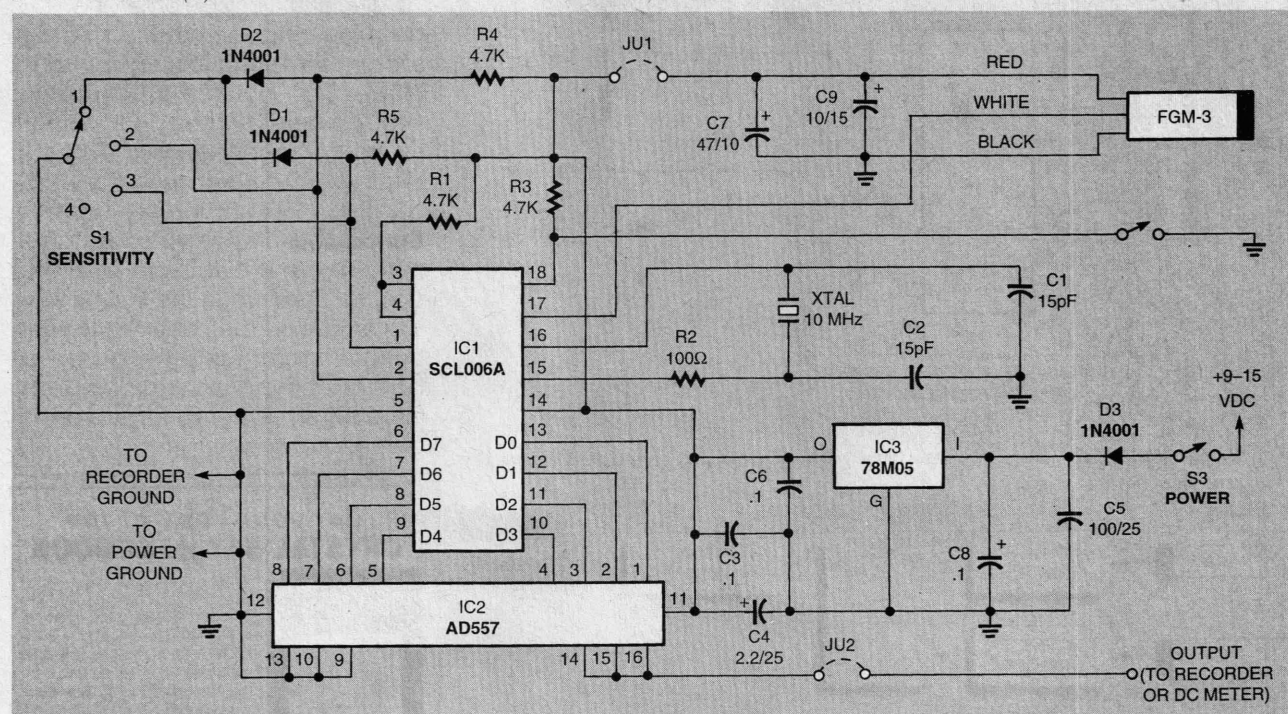


Fig. 5. Here's the circuit for a practical magnetometer. Most of the components needed and a PC board can be obtained from the source mentioned in the article.

face, or a shipwreck buried deep in the ocean silt. Archeologists use gradiometers to find artifacts and identify sites. Also, people who explore Civil War battlefields, western mining camps, and other sites often use gradiometers to facilitate their work.

Figure 6 shows the construction details for a simple gradiometer based on the FGM-3 device. It is

built using a length of PVC pipe. One sensor is permanently mounted at one end of the pipe, using any sort of appropriate non-magnetic packing material. In one experiment, I used standard 0.5-inch adhesive-backed window-sealing tape, which is used in colder areas of the country to keep the howling winds out of the house in wintertime. It worked nicely to hold

the permanent sensor in place.

The other sensor is mounted in the opposite end of the tube using an O-ring that fits snugly into the tube. Four positioning screws made of non-magnetic materials are used to align the sensor. The position of the sensor is adjusted experimentally. The idea is to position the sensor such that the gradiometer can be rotated freely in space without causing an output variation.

The gradiometer sensor is usually held vertically such that the end with the wires coming out of the FGM-3 devices is pointed downwards. That alignment allows you to find buried magnetic objects even if they are quite small.

A practical gradiometer can be built using a special interface chip by Speake, the SCL007 device. It is an 18-pin device that accepts the inputs from the two sensors, and produces an eight-bit digital output. The circuit is shown in Fig. 7. It can receive the signals from the sensors in Fig. 6, and produce a digital output proportional to the field gradient. If you want a DC output, the same sort of D/A converter used in the magnetometer of Fig. 5 can also be used for the gradiometer.

Figure 8 shows a method of using digital heterodyning to make a very sensitive gradiometer at low cost. The outputs of the two FGM-3 sensors are fed to the D input and clock (CLK) input of a D-type flip-flop. The output of the flip-flop is fed to an F/V converter such as the LM2917 device discussed earlier.

Conclusion. As we've seen, magnetic sensor projects are relatively easy to build when the FGM-3 sensor is used. The device is well behaved, and will serve nicely for both hobbyist and professional instruments, as well as for classroom demonstrations. Ω

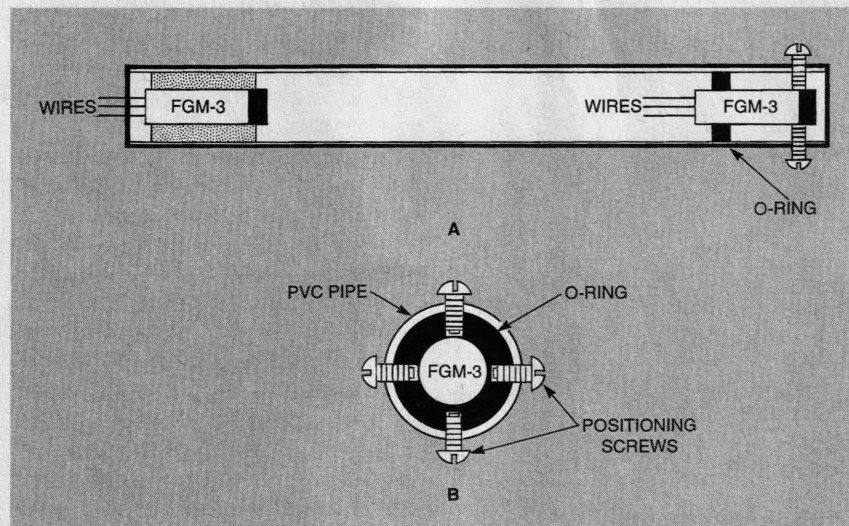


Fig. 6. This two-sensor rig forms the heart of a practical gradiometer. It is easy to build and provides excellent results.

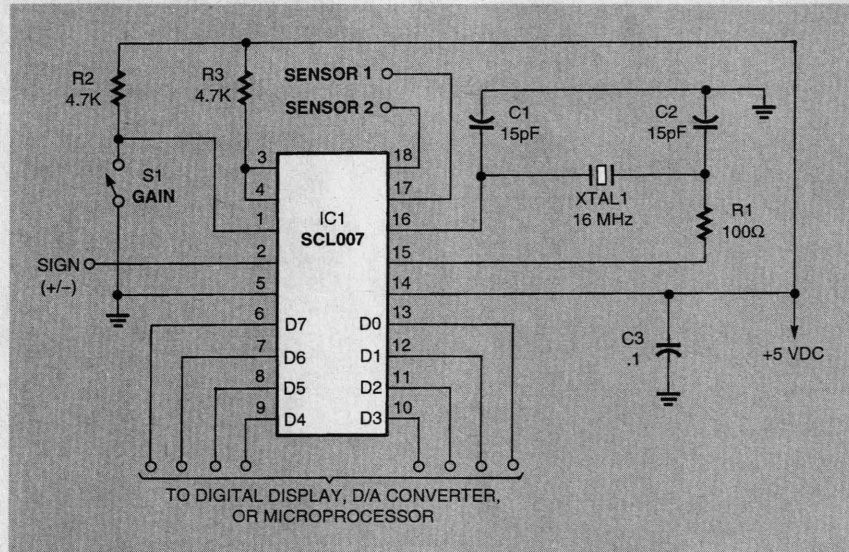


Fig. 7. Here's a circuit that converts the output of the gradiometer's sensors into a digital signal that is proportional to the field gradient.

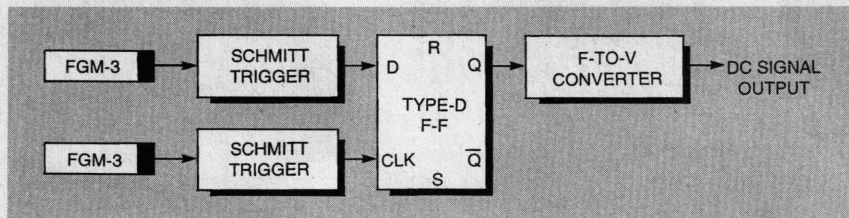
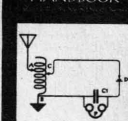


Fig. 8. Using the digital heterodyning concept discussed earlier, it is easy to build a low-cost but very sensitive gradiometer.

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THE REAL McTUBE



FRED NACHBAUR

With the phenomenon of the "retro" movement in music electronics, there has been renewed interest in the vacuum tube. Several manufacturers are again building tube amplifiers and other devices, with claims of "vintage sound" and "warm-tube distortion."

But are vacuum-tube-based amplifiers really "better" than solid-state units? That is an argument that might never be resolved; the "sound" of a particular piece of equipment is as much the listener's perception as the specifications of the amplifier circuit.

The only sure-fire way to find out if tube-based gear is right for you is to try one out. With the Real McTube device presented here, you can experiment with these interesting devices by building a tube pre-amplifier and distortion unit. The overall cost of the project will

*Give your music a real
"retro" sound with this
vacuum-tube based
preamplifier.*

depend on how many parts you already have in your junk box, but it shouldn't be over about \$50 if you shop carefully. The only part you might have trouble finding is the vacuum tube itself (along with the 9-pin miniature tube socket to go with it). Although vacuum tubes are again being manufactured in quantity (mainly in Russia), they haven't quite made it back onto the shelves of most electronics shops. If you are stuck, one source might be a musical-instrument-repair shop that specializes in vacuum-tube gear. There are also a number of companies that sell these mail order, and a number have a presence on the

Internet. Some of these are STF Electronics, 171 Springlake Dr, Spartanburg, SC 29302, Tel: 864-573-6677; SND Tube Sales, 5389 Ville Rosa La, Hazelwood, MO, Tel: 314-770-0119; Tube World, Inc., 2717 Superior Av, Sheboygan, WI 53081.

The Tube Sound. So what is it about tubes that make them sound different from solid-state devices like transistors? There is really nothing mystical about it; in reality, their unique sound derives more from their shortcomings than from anything else. Being hot, bulky and expensive, manufacturers couldn't afford to make amplifiers with massive gain, which would have let them use large amounts of negative feedback to linearize the amplifier's response and reduce distortion. The exception was for applications such as studio monitors and similar high-end equipment. You only have to lis-

ten to such tube classics as Dave Brubeck's "Take Five" or Heart's "Dreamboat Annie" albums to hear for yourself that tube equipment can be made to sound as good as top-end solid-state gear.

For musical-instrument amplifiers, however, the inherent non-linearity and high degree of even-harmonic distortion of a "raw" tube amplifier with little or no negative feedback adds to the charm of its sound. "Warm" and "gutsy" are but two of the many adjectives used to describe the sound of a vacuum-tube amplifier.

That even-harmonic distortion is especially noticeable when a tube is overdriven. Unlike a transistor, which remains reasonably linear until it reaches cutoff (no current through the device) or saturation (maximum current through the device), a tube will exhibit a softer, more gradual bottoming-out at either extreme. That form of distortion is especially pronounced at the saturation end, that is, as the voltage across the tube approaches zero. There is no sharp saturation "knee", and any attempt to define the point at which a tube reaches saturation is strictly arbitrary.

There are wide variances in those characteristics between tubes of a

given family, and even between individual tubes of a given type. Additionally, a tube's characteristics change as the tube ages. That can be a real nightmare if you're trying to design a consistent and predictable piece of gear using tubes. However, those factors all add up to the mystique of the vacuum tube for musicians and experimenters. I've heard of musicians who treasure old, worn-out, gassy 12AX7 amplifier tubes because of the particularly dirty distortion that they're capable of.

The 12AX7 is one of those "classic" tubes that was used in virtually all vintage tube gear, mainly because of its relatively high gain and reasonable linearity. As tubes go, it is relatively quiet—notwithstanding the high degree of thermal noise inherent in all vacuum tubes. The 12AX7 (and its European equivalent the ECC83) is at least mechanically constructed to minimize "microphonics," noise caused by mechanical vibration of its internal elements.

The 12AX7 is but one of a whole family of dual-triodes tubes. Other devices in the family are the 12AT7, 12AU7, 12AY7, 12AZ7, and numerous European types. The filament (heater) requirements are all the

same—either 12.6 volts at 150 mA as suggested by the first two digits of the type numbers or 6.3 volts at 300 mA depending on how the two filaments are connected. Another boon is that the pinouts for all of those types of tubes are identical. The connections are (going clockwise from pin 1): Plate, Grid, Cathode (triode 1), Filament, Filament, Plate, Grid Cathode (triode 2), and Filament Center. The Real McTube will work with any of the tubes in that family, giving you the opportunity to experiment with different types to get the exact sound you're looking for.

Basic Tube Theory. Unlike transistors, tubes require at least two power supplies. Before the advent of universal home power, tube-based electronic gear was operated on batteries. One battery (called the "A" battery) was needed for the low-voltage, relatively high-current supply to heat the tube's filament. Another battery (the "B" battery) was the high-voltage, low-current battery used to power the tube's plate circuit. That labeling convention has persisted to this day; if you ever wondered why you hear of devices connected to "B+", now you know. In the

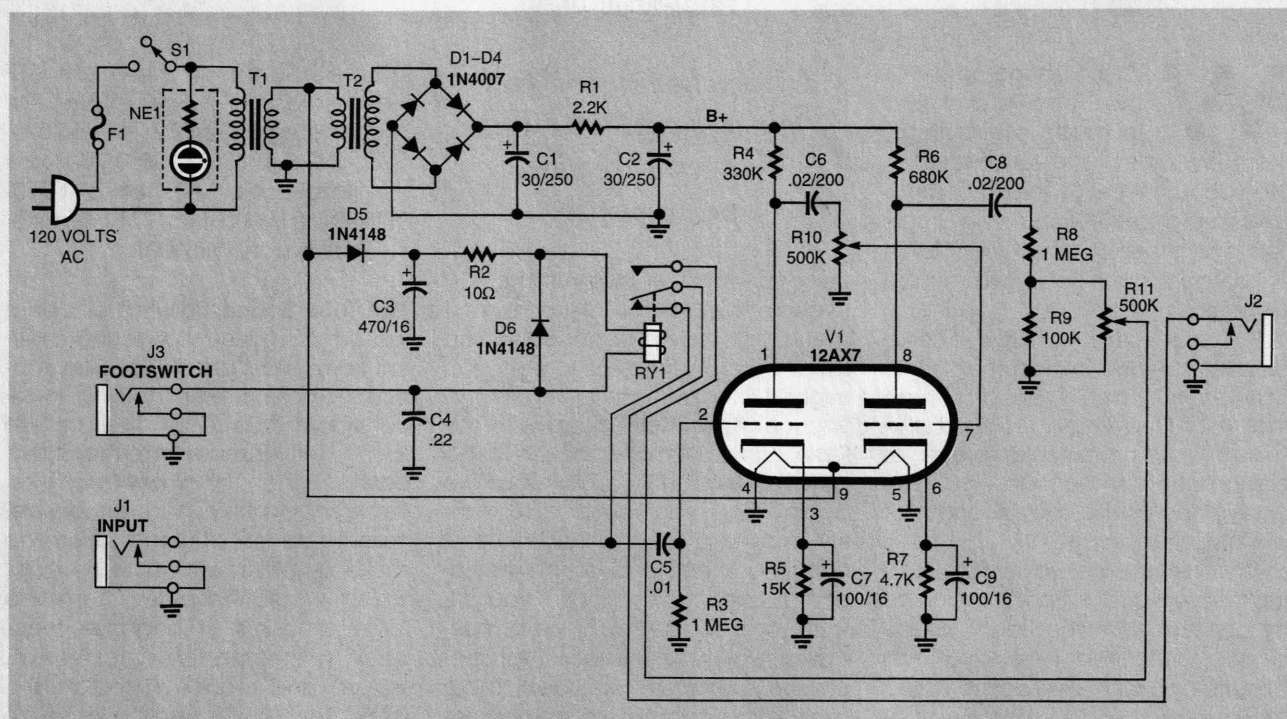


Fig. 1. The Real McTube is simply a two-stage, non-inverting preamplifier. Using a vacuum tube for the active element can create many interesting sounds—something musicians are always looking for!

early days, there was even a "C" battery. It was used to provide the negative-grid bias for some circuits. That need didn't continue for very long except in special radio applications because the negative-grid bias can usually be obtained by using a cathode resistor; that trick is used in the Real McTube.

Tubes are roughly analogous to field-effect transistors (FETs) in that they are *voltage amplifiers*; bipolar transistors amplify current. The tube's cathode is like the FET's source, the grid is the gate, and the plate is the drain. Like FETs, tubes have a near-infinite input impedance, and a high output impedance. The most widely-used tube configuration—the common-cathode arrangement—is quite similar to the common-source FET-amplifier configuration. But the similarity ends at that point.

Physically, a tube operates by heating the cathode, one of the electrodes inside the evacuated glass envelope. That causes electrons in the cathode to attain enough velocity to leave the surface, forming a "space charge" around the cathode. The positively-charged plate (also called the *anode*) surrounds the cathode with the cathode being at the center. The plate attracts the space charge, causing an electron current to flow through the space between the cathode and plate. If the plate is negatively charged, the space charge is repelled and no current flows; that is how the vacuum diode works.

In what's called a *vacuum triode*, there is another electrode called the *grid* that is placed between the cathode and the plate. It usually consists of a spiral screen of fine wire. The grid is normally biased negatively so that it does not act as a secondary plate and draw current. Small changes in voltage on the grid cause significantly greater changes in electron current through the tube from cathode to plate, resulting in amplification.

As the vacuum tube was developed, additional grids were added to form first the *tetode*, then the *pentode*. The culmination of multi-grid mania was the *heptode*, which is a five-grid device that was once used extensively as the oscilla-

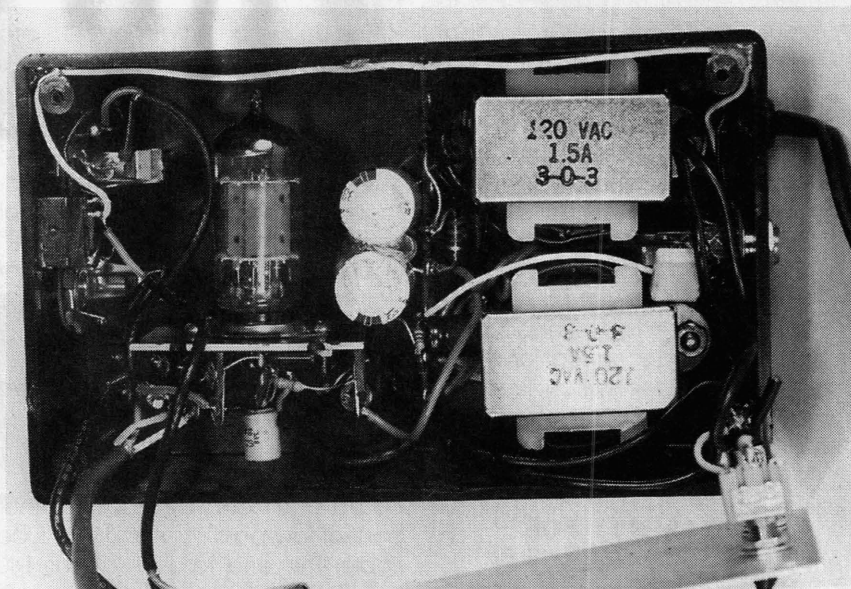


Fig. 2. Here is the author's prototype for the Real McTube. While a smaller case makes for a portable unit, assembly can be difficult.

tor/mixer in home radio receivers.

When dealing with vacuum tubes, it makes sense to consider current flowing from the negative terminal to the positive terminal of the "B" supply because that's what physically happens. That is called the *electron-current* convention, and was taught in most technologist courses up until a couple of decades ago. The problem with that convention is that the algebraic signs end up backwards in circuit design and analysis. Engineering courses thankfully used the opposite convention, called *conventional-current* flow, in which current is assumed to flow from positive to negative. That has the advantage in semiconductor circuits of making the arrows in diode and transistor symbols point in the right direction. When dealing with tubes, however, which convention you use is strictly a matter of personal preference.

Finally, it should be pointed out that unlike transistors, tubes come in only one polarity because we only have one type of charge carrier—electrons. Unlike solid-state technology, no one has yet to figure out how to make holes in a vacuum. I suppose it's theoretically possible to make tubes out of antimatter in order to get "PNP" tubes, but storage and interfacing could be a bit of a headache. Such devices will probably remain in the realm of science fiction.

How It Works. The Real McTube is a simple circuit; it can be seen in Fig. 1. Line current is applied through F1 and S1 to the primary of T1. A neon bulb with an integral resistor, NE1, indicates whenever the Real McTube is turned on. The secondary of T1 provides the 6.3 volts needed to power the tube's heater filaments. The tube filaments are connected in series with a center tap. Since we are supplying 6.3 volts, power is connected to the center tap; the other sides of the filaments are grounded. The tube is set up that way in order to use either a 6.3-volt or a 12.6-volt "A battery". With a 6.3-volt source, the filaments are effectively wired in parallel. If T1 were a 12.6-volt transformer, the two end terminals of the heaters would be used, leaving the center tap unconnected; that would put the heaters in a series circuit.

The output of T1 also provides voltage to bypass relay RY1. Diode D5 half-wave rectifies the voltage, and C3 filters and smoothes the pulsating DC in order to prevent noise and relay chattering. Resistor R2 limits the current draw by RY1 and provides a small voltage drop for the relay. Diode D6 absorbs any voltage spikes that are generated when RY1 turns off. Capacitor C4 helps to eliminate any transients that might cause audible clicks when RY1 changes state.

Jack J3 is wired in such a way that when nothing is plugged into it, RY1 is engaged whenever the unit is turned on. That way, the input signal from J1 is routed through the pre-amp circuit to J2. If the unit is turned off, the relay disengages and the signal is routed straight through. A simple single-pole, single-throw footswitch is plugged into J3; it lets the Real McTube be used or bypassed as needed without having to disconnect or move any wires. The footswitch is simply wired to short the contacts of J3 in order to operate RY1.

PARTS LIST FOR THE REAL MCTUBE

SEMICONDUCTORS

D1-D4—1N4007 silicon diode
D5, D6—1N4148 silicon diode

RESISTORS

(All resistors are 1/4-watt, 5% metal-film units unless otherwise indicated.)

R1—2200-ohm
R2—10-ohm
R3, R8—1-megohm
R4—330,000-ohm, 1/2-watt
R5—15,000-ohm
R6—680,000-ohm
R7—4700-ohm
R9—100,000-ohm
R10, R11—500,000-ohm panel-mount potentiometers, audio taper

CAPACITORS

C1, C2—30- μ F, 250-WVDC, electrolytic
C3—470- μ F, 16-WVDC, electrolytic
C4—0.22- μ F, ceramic-disc
C5—0.01- μ F, 100-WVDC, ceramic-disc
C6, C8—0.02- μ F, 200-WVDC, Mylar
C7, C9—100- μ F, 16-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

F1—0.25A, 125-volt AC fuse, slow-blow
J1, J3—1/4-inch mono switching jack
J2—1/4-inch mono jack
NE1—Neon indicator, 125-volt AC, with built-in resistor
RY1—Single-pole, double-throw relay, 6-volt DC coil
S1—Single-pole, single-throw switch, 125-volt AC contacts
T1, T2—6.3-volt, 1.5-amp filament transformer
V1—12AX7 dual-triode vacuum tube
Power cord and plug, strain relief, enclosure, 9-pin miniature chassis-mount tube socket, SPST footswitch, mounting bracket for tube socket, knobs, hardware, terminal strips, wire, shielded cable, etc.

The output from T1 is coupled into an identical transformer, T2, that is wired "backwards." That steps the voltage back up to about 110 VAC for the plate supply, or "B battery." That arrangement has the advantage of providing line isolation for the Real McTube, reducing the chance of short circuits or shocks of damaging any equipment that is plugged into the Real McTube—including the operator!

Diodes D1-D4 form a full-wave bridge rectifier for the B+ voltage. Capacitor C1 filters the resulting pulsating DC, and the network composed of R1 and C2 adds another pole of low-pass filtration to reduce ripple. The end result is about 140 volts DC under load to power the plate circuits of the Real McTube.

An audio signal from J1 is coupled to the grid of the first stage through C5. Input resistor R3 serves a dual function. Primarily, it acts as a "grid leak" to prevent any electrons that accumulate on the grid from piling up on C5; the resulting negative voltage would eventually cause the tube to approach cut-off. Secondly, it serves to bias the grid at very near ground potential.

The plate-load resistor is R4. The varying plate current flowing through R4 causes a voltage drop that is proportional to the change in current. Therefore, R4 plays a major part in defining the voltage gain of the amplifier stage.

The plate current also flows through cathode resistor R5, causing a smaller voltage drop across it. The result is that the cathode is positive with respect to ground. Since the grid is at (or close to) ground potential, it follows that the grid will be negative with respect to the cathode. That is how we get the necessary negative-grid bias without having to resort to a third power supply—the "C" battery. Capacitor C7 "shorts out" any AC signal component, eliminating the negative feedback that would otherwise result. That has two major effects: first, it maximizes the voltage gain, making it easier to achieve overdrive; and secondly, it eliminates any reduction in non-linearity distortion that would defeat the purpose behind the Real McTube—getting "that tube sound."

The signal output at the plate of the first stage is coupled into a similar stage by capacitor C6 and gain control R10. The component values in the two stages were chosen more or less experimentally to suit the author's personal tastes.

The output of the second stage is routed through an attenuator consisting of R8, R9, and output control R11. The accumulated gain after the second stage is so high that the absence of an attenuator would make it ridiculously difficult, if not impossible, to adjust the controls for varying degrees of overdrive. The wiper of R11 is the output of the Real McTube. The final signal passes through the contacts of RY1 to output jack J2.

Building the Real McTube. The prototype for the Real McTube was built in a plastic box with an aluminum lid for the controls; the size of the unit was large enough to hold all of the components yet small enough to be portable. Admittedly, the resulting layout is a little tight—the layout of the author's prototype is shown in Fig. 2.

As in most traditional vacuum-tube units, there is no PC board for the Real McTube; all of the components are mounted onto terminal strips and connected directly to each other. The tube socket was mounted on an L-shaped metal bracket that measures about 2-inches square with terminal strips mounted to the same screws that are used to hold the socket to the bracket. The various biasing and coupling components are then soldered between the socket and the terminal strips.

The power supply was built in the case itself, again using terminal strips to hold all of the components except the transformers, S1, and NE1. The input and output jacks were installed on the front of the box, and the remote footswitch jack on the rear.

You might want to leave yourself a bit more space for experimenting—if so, you should obviously use a larger box. With a single 12AX7, heat is not a big consideration so you don't have to worry about ventilation. Orientation of the tube doesn't matter either; the 12AX7 family can

operate in any orientation—which is not necessarily a given for all tubes!

Be careful about wiring, especially in the comparatively high-voltage plate circuits. Be sure to use capacitors that will stand the voltage; most standard ceramic-disc capacitors are only rated for 25 or 50 volts. Having a capacitor explode into flames a few minutes after powering up the Real McTube for the first time is definitely an experience that you'd like to avoid at all costs! On a similar note, it is also a good idea to stay away from carbon-composition resistors, especially for R4 and R6, or you can get another classic tube sound—lots of hiss.

All of the signal lines, such as the lines to J1, J2, and RY1 should be wired with good quality shielded cable. A good choice would be the RG174 variety. Keep the component leads as short as possible by mounting them as close to their associated tube socket pins as is practical. The input to the first stage is especially critical; we're dealing with a very high input impedance and low signal level. Hum and noise pickup is a real concern under those conditions.

Some components in the power supply can be substituted without affecting the operation of the Real McTube; if you have an item on hand, it can help keep the cost of the project down. For example, the 6-volt transformers can be replaced with 12-volt units. If you do that, the tube's heaters will need to be wired in series instead of in parallel as shown. Simply connect the supply to pins 4 and 5; leave pin 9 unconnected. A 12-volt relay will also have to be used for RY1. If you want to use 12-volt transformers but have a 6-volt relay, increase the value of R2 to compensate. That is done by measuring the DC resistance of the

relay coil and choosing the next higher standard value for R2. For example, if the resistance of your relay is 50 ohms, use a 56-ohm resistor for R2.

Diodes D1-D4 can be substituted with a single bridge rectifier.

One important safety consideration: **DO NOT** be tempted to omit the second transformer! Although the circuit will work if you derive the 120 VAC directly from the line, you can seriously injure or even kill yourself or somebody else. At best, you can ruin your gear if you don't isolate the high-voltage supply from the AC line.

Testing the Real McTube. Once the circuit is wired up, stop and check your work for errors or accidental shorts. Once you are satisfied with your workmanship, plug the unit into a wall socket and turn it on *without* V1 being plugged into its socket. With a voltmeter, measure the voltages at pins 1 and 6 of the tube socket; you should read about 145 volts DC. At pin 9, you should read about 7 volts AC. At all of the other pins, you should read 0 volts.

If all is well so far, unplug the unit and insert the tube into the socket. You should be careful not to accidentally touch the filter capacitors—they will carry a residual charge under those conditions. A good safety tip is to short them to ground with a 1000-ohm resistor for several seconds whenever you work on the unit after having powered it up. While monitoring the B+ voltage at the positive end of C2, turn the unit on. You should see the heaters in the tube start to glow. After about 10 seconds, the B+ voltage should sag a bit as the triodes start conducting. The voltage should drop no lower than about 135–140 volts DC.

Measure the voltages on the pins

of the tube again. If you used the component values suggested, your readings should be close to the values shown in Table 1. If the plate voltages are excessively low or if one or both of the grids are positive with respect to ground, power down right away and find the problem. The same advice applies if you see blue flashes in the tube or smell anything unusual. Re-examine your work and double-check the values of the components. Be sure that you are using good quality Mylar or similar capacitors for C6 and C8, and that they have at least a 150-volt rating.

Assuming that the voltages are correct, turn the unit off. Connect the output to an amplifier and the input to your instrument. Turn the gain and output controls to minimum, and power up. Slowly bring up the controls until you hear your instrument through your amplifier.

With the Real McTube working, you can now start using and experimenting with it.

Using the Real McTube. What sound you get from your guitar or other instrument with the Real McTube will depend greatly on the settings of the various controls—specifically, the volume and tone controls on your instrument and the Real McTube's gain control. The output control has little or no effect on the sound, and is used primarily to balance the volume of the sound through the unit.

Note that both stages are inverting amplifiers; the overall phase of your signal will therefore not change. However, the adjustments of your controls will determine what portion of your signal will be clipped. The first stage will tend to clip more on negative half-cycles as the tube reaches cutoff—remember the discussion above regarding cutoff versus saturation. The second stage, however, will tend to clip more on positive half-cycles, since it sees an inverted signal at its input. By carefully balancing your instrument controls and the gain control, virtually any distortion sound from a mild "warm-fuzzy-feeling" to a hard "monster-metal-shred-head" sound, with lots of odd-harmonic distortion, can be attained.

A compressor before the Real

(continued on page 50)

**TABLE 1—
VOLTAGE MEASUREMENTS AT V1**

Pin	Description	Voltage
1	Plate A	+108 VDC
2	Grid A	–0.001 VDC
3	Cathode A	+1.5 VDC
4,5	Filaments	0 volts
6	Plate B	+60 VDC
7	Grid B	–0.1 VDC
8	Cathode B	+0.7 VDC
9	Filament	6.3 VAC

In the beginning there was coherent light

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THIS NEW ONGOING COLUMN BEGINS BY PRESENTING A BRIEF HISTORY OF LASERS, SOME BASIC THEORY, AND DETAILS OF HOW TO BUILD A LASER YOU CAN USE TO CONDUCT A WIDE VARIETY OF EXPERIMENTS THAT

will be presented in future columns.

In the early 1900's, Albert Einstein published a paper addressing the properties of what he called stimulated emission of radiation or SER. His theory explored the principals of photon production through inner atomic activity. It solidly laid the groundwork for what is obviously one of the most important inventions of this century; the "LASER" the anacronom for light amplification by stimulated emission of radiation.

It wasn't until the 1950's that Bell Laboratories began to search for laser applications. Earlier research was hampered by financial restrictions, the demanding needs of the military during two world wars, and the perception by many that the final result would amount to little more than a fancy flashlight. In the end, however, all the development frustration, time, and money expended to produce that first blast of laser red light, from a rod of synthetic ruby crystal, was worth the wait.

In the forty years since its introduction, lasers of all types and varieties have become an important part of our lives. They are used by surveyors to measure distances accurately, in manufacturing plants

to control the smoothness of surfaces with great precision, and they even show up in such everyday places such as automobile speed traps, CD and videodisc players, and electronic pointers used by lecturers.

Laser light theory and properties

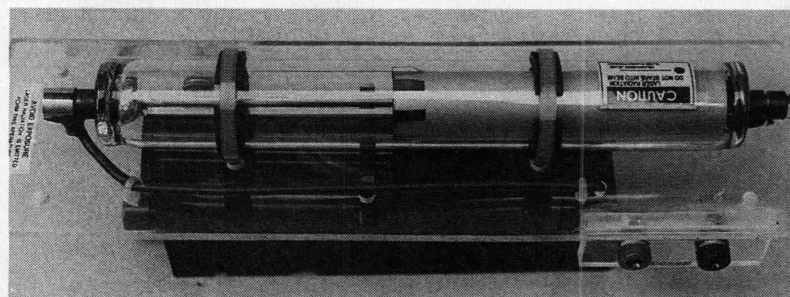


FIG. 1—LASER TUBE CLOSE UP. This one is the Spectra-Physics model No. 088. It is shown mounted on a Plexiglass platform with the high-voltage power supply directly below it.

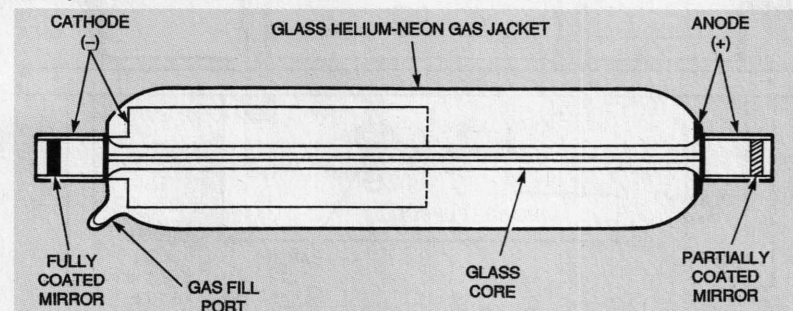


FIG. 2—BASIC DIAGRAM OF AN He-Ne laser tube. Some tubes have metal ends instead of the solid glass envelope. The glass fill port is used during manufacturing to charge the tube.

WARNING!!!

This article deals with and involves subject matter and the use of materials and substances that may be hazardous to health and life. Do not attempt to implement or use the information contained herein unless you are experienced and skilled with respect to such subject matter, materials and substances. Neither the publisher nor the author make any representations as for the completeness or the accuracy of the information contained herein and disclaim any liability for damages or injuries, whether caused by or arising from the lack of completeness, inaccuracies of the information, misinterpretations of the directions, misapplication of the information or otherwise.

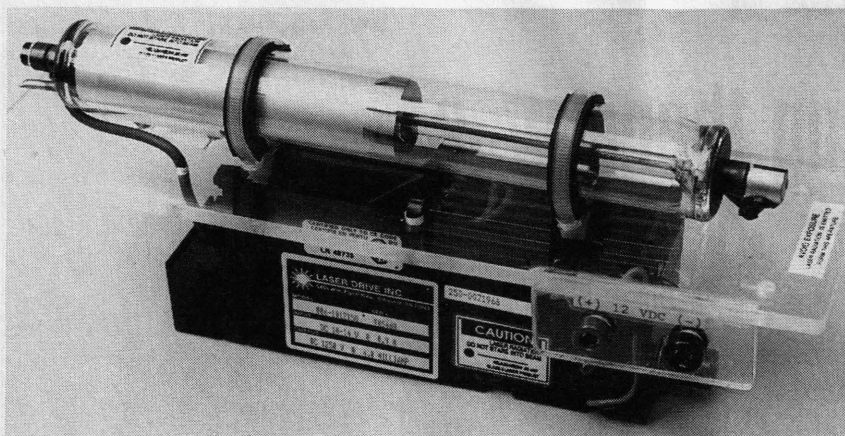


FIG. 3— LASER TUBE MOUNTED ON ITS PLATFORM. This side view shows details of the mounting techniques that are used.

In theory many substances can be made to lase if the proper energy level of stimulation is applied. There are three major categories of lasers (those made from solid, liquid, and gaseous materials. There are also four other lesser-known groups (diode lasers such as the ones used in your compact disc player; tunable dye lasers that can emit laser light at more than one wavelength; X-ray lasers that do not produce a visible light; and a device known as a free-electron laser that uses a beam of free electrons (electrons not associated with any atom) as the active material.

The energy source that activates a

laser can be light, heat, electricity, or even atomic energy. Solid lasers are built around rods of ruby, or other, synthetic gem materials. Other solid lasers use materials such as Nd-YAG (YAG stands for yttrium-aluminum-garnet. Nd-YAG is a YAG rod doped with neodymium) and Nd-glass (glass doped with neodymium) materials. Semiconductor lasers also use solid materials but work in an entirely different manner. We will not look at these devices in this series.

Liquid-based lasers use both dyes and chemicals, that are usually pumped through the lasing chamber where they are exposed to the energy

source. Perhaps the most commonly used medium is gas, if we don't count the tens of millions of relatively low-power lasers used in everyday consumer products such as CD players, CD-ROM drives and other devices of this kind. Helium, neon, argon, krypton, carbon dioxide, and even some vaporized metals, such as cadmium or arsenic, are also used in conjunction with other gases as laser materials.

Whether solid, liquid, or gas, a material will lase if more of its atoms are in a high-energy state than in a rest state. This doesn't normally occur in nature, but must be forced, or in the terminology of lasers, pumped.

Until Einstein proposed his idea of SER, scientists believed that a photon could only be absorbed by an atom (and raise that atom to a higher energy level) or be emitted by the atom (as it dropped to a lower energy level). Einstein was most interested in what would happen if a photon hit an atom that already was at a higher energy level. He theorized that a photon with energy corresponding to that of an energy-level transition could stimulate the release of a photon that would be identical to the first. If enough atoms could be excited, the chance of photons hitting them would be increased. That would lead to a chain reaction where photons would hit excited atoms and make new photons, with the process continuing until the energy source was removed.

Normally, stimulated emission is unlikely because when a material is in equilibrium, more atoms are in low-energy states than are excited, and the population at a particular energy state decreases as the energy of the state increases. In other words, atoms tend to drop to the lowest available energy level by spontaneously emitting a photon.

For stimulated emission to occur, a population inversion is required. In other words the population of atoms at a higher-level energy level must be higher than that at a lower-energy level. When that happens, photons that are emitted spontaneously are more likely to collide with an atom in an excited state than to be absorbed by the atom with a lower energy level. The result is laser gain or light ampli-

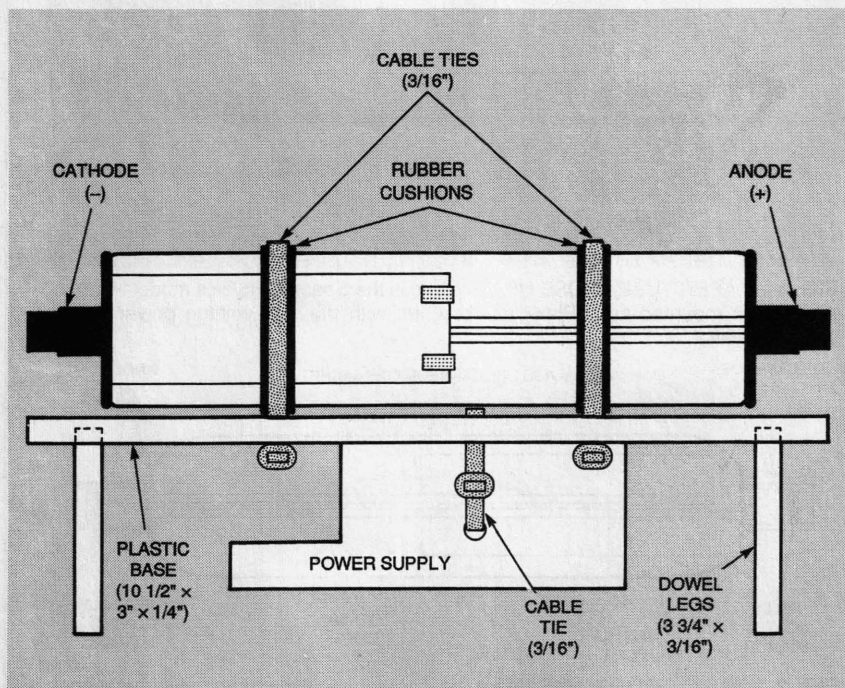


FIG. 4— ANOTHER VIEW OF HOW THE LASER TUBE is fastened to its platform. For safety reasons, it is best to place the laser tube inside a plastic or metal housing.

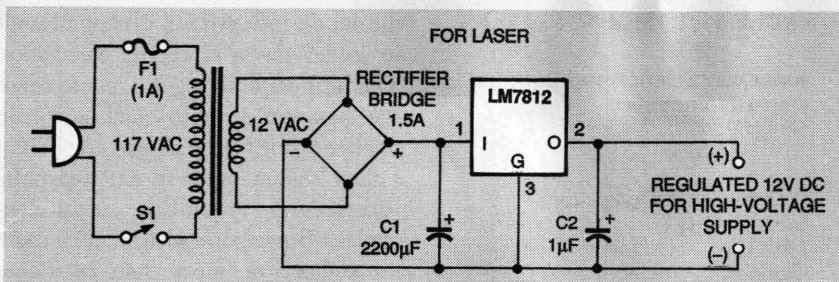


FIG. 5— REGULATED POWER SUPPLY DELIVERS 12 volts DC to drive the laser's high-voltage supply.

fications as the chain reaction creates more atoms with higher energy levels.

Stimulated emission has a couple of special properties that make laser light unique (it has the same wavelength as the original photon (monochromatic), and it is in-phase (or coherent) with the original light. A third special property of laser light is collimation (the laser emits a narrow beam of light that exhibits very little spreading, even over long distances. Collimation occurs without using any optics, and is responsible for the high intensity of the reflected laser light or "spot". Semiconductor lasers emit beams with a divergence angle much higher than those of gas or solid lasers. Optics are used to shape the beam to make it narrower.

The most commonly used laser energy sources are light and electricity. With solid and liquid lasers, light, (in the form of high-power xenon flash tubes or carbon arc lamps arranged to surround the laser tube (is

gaseous lasers, an envelope, often made of glass, contains the gas, or combination of gases, to be lased. A core running the length of the envelope, through its center, terminates into a mirror-electrode assembly at each end. One end is the anode. The other is the cathode. A DC voltage is connected; positive to the anode and negative to the cathode. This voltage, usually between 300 and 10,000 volts, depending on the size and type of laser, provides the energy for the population inversion. While its efficiency is a mere 2 to 5 %, the value and versatility of device far outweighs the cost of operation.

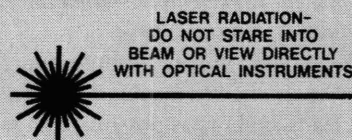
The power supplies used are also special. Most of the time a very high "starter" voltage, about 10,000 volts, is required to initiate the population inversion. Then the supply drops back to somewhere between 1000 and 4000 volts DC, to maintain that state.

Up to this point, most of the information has been directed at the stimu-

lope must be precisely parallel to each other, and at a 90-degree angle to the sides of the device. Each end is fitted with a mirror that reflects the photons back and forth, and produces the amplification. As more photons are emitted, they combine with the ones already present.

If the system was completely light tight, the amplification process would continue to saturation and there would be no laser beam output. To permit the laser to emit its beam, the mirror on one end, usually the anode, is only a partial mirror. This allows

CAUTION



CLASS II LASER PRODUCT
LESS THAN 1-mW OUTPUT

FIG. 7— THIS IS A WARNING NOTICE THAT MUST be attached to all low-power lasers. Make sure you attach one to your unit.

some light to escape. The quantity of light that escapes varies to some degree among various manufacturers, but the rule of thumb is about 10%. The escaping light is what becomes the intense laser beam.

Build your own laser

With the technical stuff out of the way, it's time to get down to the fun. There are numerous sources of laser parts, power supplies, equipment and support items, such as front surface mirrors, motors, and optical elements. Table 1 lists some of these suppliers. Many of these companies carry a wide variety of types in all three states. We are going to show you how to build a small low-power helium-neon, He-Ne gas laser. Helium-neon tubes are relatively inexpensive and they are readily available, along with power supplies from surplus-equipment suppliers.

He-Ne lasers are remarkably reliable. It is rare to find a sub-performing tube that works only partially.

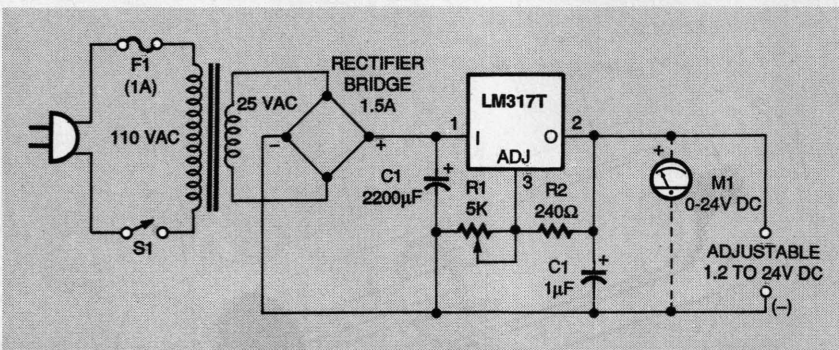


FIG. 6— A MORE ELABORATE VARIABLE POWER SUPPLY is shown here. The meter, which can be an analog or a digital type, is optional.

a practical and effective choice. This type of laser is usually a pulsed emitter that sends out short, high-power bursts of light rather than a continuous beam. The high-intensity flashes of light provide the energy to create the population inversion. With

lated emission of radiation (SER) aspect of the laser. But the light amplification (LA) part is equally important. Let's take a look at how lasers go about this procedure. Whether solid, liquid, or gaseous; the ends of the rod or the liquid/gas enve-

The tubes almost always either work very well or not at all. Most suppliers are quite good about testing used equipment and offer limited warranties to protect their customers. Surplus lasers can provide excellent savings for the experimenter. The average He-Ne laser has a life expectancy that exceeds 5,000 hours.

The second and very important consideration is **SAFETY!** This is a good time to mention the hazards of working with lasers and the precautions that must be followed. Remember, the light emitted is very intense. *Take no chances. Never look directly into the beam of any laser. Never allow the beam to reflect back into your, or anybody else's eyes. Permanent damage to the retina can easily occur.*

The high-voltage supply for the tube is the next consideration. Keep the leads as short as possible, and well insulated. Keep the connections, on

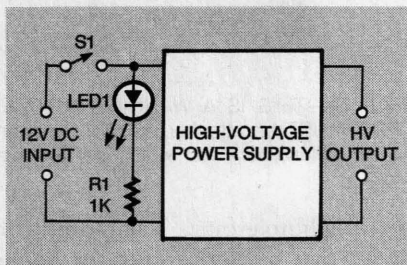


FIG. 8— DETAILS OF THE PROTECTIVE HOUSING for the laser tube.

both ends, out of easy touch. The average supply will have a "kicker" voltage of between 6,000 and 15,000 volts, and will run continuously at 1,000 to 3,000 volts DC, depending on its power. Even at the low current in the 4 to 10-mA range, contacting a hot He-Ne power supply can be dangerous. With this in mind, let's take a first look at the unit we will build.

The laser tube we used is a Spectra-Physics No.088, 1-mW helium-neon device that measures 9 3/4 inches long by 1 1/4 inches in diameter. Its electrodes are at each end; the cathode is attached to the long metal sleeve that surrounds the glass core. You can clearly see the details in the photo in Fig. 1 and the diagram in Fig. 2. The power supply's negative output (Laser Drive Inc. No. 006-1012158) is connected to the cathode, while the posi-

TABLE 1—LASER SUPPLIERS

American Science and Surplus

3605 Howard Street
Skokie, IL 60076
(708) 982-0870

Electronics & Computers Surplus City

1490 West Artesia Blvd.
Gardena, CA 90248
(310) 217-1922

Electronic Rainbow, Inc.

6254 LaPas Trail
Indianapolis, IN 46268
(317) 291-7262

Information Unlimited

Box 716
Amherst, NH 03031
(603) 673-4730

Midwest Laser Products

Box 2187
Bridgeview, IL 60455
(708) 460-9595

MKW Industries

1269 Pomona Road
Corona, CA 91720
(800) 356-7714

Timeline, Inc.

23605 Telco Avenue
Torrance, CA 90505
(301) 784-5488

the anode and cathode, do not attempt to solder the wire to the electrodes. The heat of soldering is likely to cause the glass envelope to crack and ruin the laser tube. Instead, use one half of a fuse holder, or wrap and twist the wire securely around the metal electrodes. It is a good idea to keep these leads short, no more than 24 inches long, for both safety and efficiency. If you follow the recommended assembly methods, long leads to the tube are not needed. Mount the laser together with the high-voltage power supply as shown in Fig. 3. This method works well, as it keeps everything together, in one easily handled package.

The power-supply assembly is an epoxy encapsulated, portable unit, that measures 5 inches x 1 1/8 x 1 inches, in a plastic case. At one end, a length of color coded (red for positive, and black/yellow for negative) small-gauge wire protrudes for connection to the 12-volt DC, 1 ampere, standard power supply. On top, there are two spring-clip quick connection terminals (one each for the anode and cathode). The positive terminal is at the open end of the package (see Fig. 4).

Here, the high-voltage leads can be soldered. As an alternative, well insulated spade connectors also make good choices for this hook-up. Since the input to the high-voltage supply is 12 volts DC at 1 ampere, the laser can be made portable by using a battery pack

When making the connections to

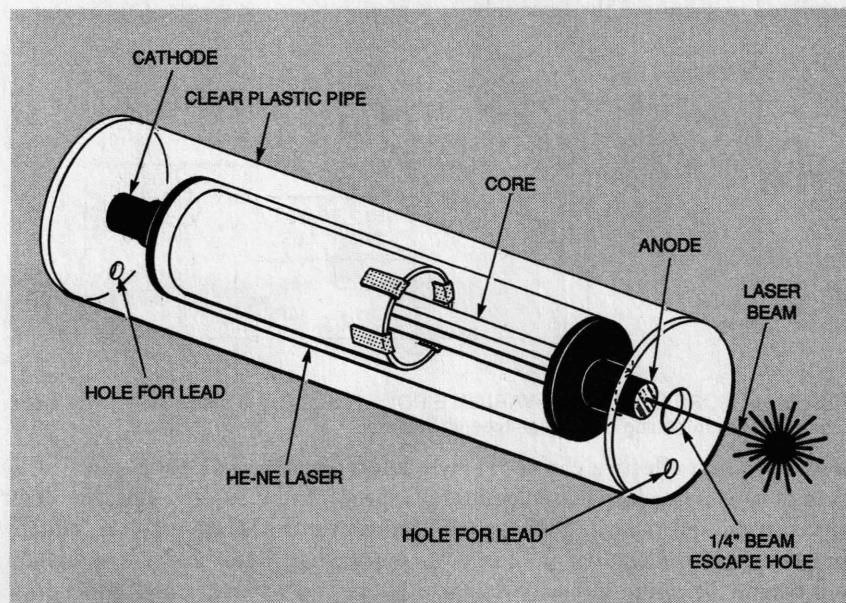


FIG. 9— ON-OFF WCH AND INDICATOR LIGHT CIRCUIT. You can add it to the laser's high-voltage power supply.

for power. However, for most of the experiments this column will describe, the 12-volt regulated power supply (see Fig. 5), is all you will need. Once you have the wiring ready, connect the three components together temporarily and apply AC power to the 12-volt supply for a systems check. If all is well, the tube will energize and emit a red beam of light. To protect yourself from any electrical shock, do not contact the laser while it is operating and exposed like this.

If the test run is successful, disassemble the components and start reassembling the laser and power supply in a more permanent arrangement. Figure 5 shows one layout, but is not the only way to go. Here, a piece of Plexiglas, 1/4 inch thick by 10 1/2- by 3 inches is cut to form a platform to hold the laser and its power supply. Drill the necessary mounting holes and use cable ties as shown together with soft rubber strips as cushions to secure the tube to the board. Next, drill 3/16-inch holes half way down the platform, 3/4-inch in from the sides..

Now make the anode and cathode connections. Secure the power supply to the underside of the Plexiglas board with cable ties. Feed the anode and cathode leads to the tube through 1/8-inch holes in the board. Connect them to the laser tube to complete assembly (see Fig. 6). To comply with the regulations of the US Center for Devices and Radiological Health (C.D.R.H.), a warning label must be placed prominently on the completed assembly. Figure 6 is a facsimile of the needed label.

For added safety you can place the laser tube into a housing. There are several ways to do this. One favorite method is to use an appropriate length of clear plastic pipe with ends (see Fig. 7). This method leaves you with clear view of the laser tube and allows you to observe the lasing action (glowing of the glass core) when the unit is operating. The clear pipe can easily be masked with black paper to block the light, when you do not want the extraneous light. Whatever type of housing you choose, you need to drill

a 1/4-inch hole in the cap at the emitting end to provide an unobstructed exit for the beam. Also, don't forget to drill holes for the high voltage leads that go to the anode and cathode. These can be located at the ends of the housing. The wires can be brought back and secured with cable ties at the point where they enter the power supply. In most applications the ballast, if used, remains outside the tube housing, for heat protection.

Now that the laser assembly is complete, cement 2 1/2- by 1-inch Plexiglas strips to each side of the board. Use 1/4-inch holes to accommodate banana jacks for the 12-volt DC power supply connection and anode lead access. These will be used later during the electronic modulation experiments (see Fig. 8).

The next installment show you how to put your newly assembled laser to work. You'll learn how to modulate the beam both mechanically and electronically. You will transmit sound signals over the beam and perform other fascinating laser experiments. **EN**

HARDWARE HACKER

continued from page 48

reader. The DOS and UNIX versions apparently are not being upgraded, at least not so far. Also, the Acrobat "security" levels seem to have serious problems.

Certain uploading protocols, especially XMODEM, can distort the file just enough to prevent valid users from ever gaining access. And, if you give your user the ability to print his file, he can subvert *all* other security levels, any time he wants to. More details have been posted in the file INSECURE.TXT.

I do not personally use any of the Acrobat security features.

If you want to jump into Acrobat, their *Adobe Developer's Association* has a \$500 yearly developer program. This gives you the latest insider CD's with all of their latest software.

I've posted scads more info, tech docs, and great heaping bunches of Acrobat support to GENIE PSRT. I stock the Acrobat "pewter" reference manual here at Synergetics.

New Tech Lit

From *Texas Instruments*, a CD on *MOS Memory Specifications*. Plus a new *Power Semiconductors Product Digest* from *International Rectifier*.

The full text of all of the FCC regs are now available on the internet as <http://www.pls.com:8001/his/cfr.html>. Actually, the *entire* CFR or *Code of Federal Regulations* is stashed here. With powerful search features.

Those miniature melody generators as used in greeting cards and such are distributed in quantity by *Dicker*. A melody chip source is LSI/CSI.

Some background info on those VCR *Plus* codes has been posted to PSRT as file VCRPLUS.TXT.

Industrial Market Place is a great surplus trade journal. This is mostly mechanical stuff, but there's lots of computers and electronics as well.

WebWeek is a new net magazine by the *PCWeek* and *MACWeek* folks. Free to qualified subscribers.

Samples of porous plastics are now offered by *Porex Technologies*. The specs on low melting point alloys are in an *Indium Corporation* flyer.

A major improvement in polymer

solar cells is described in *Science* for December 15, 95. Pages 1789-91.

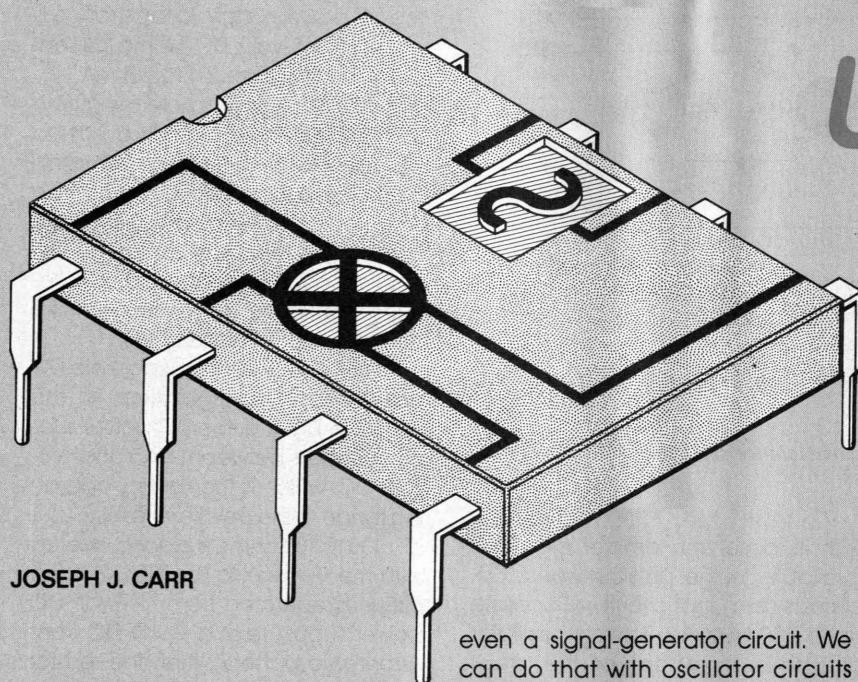
A free coaxial cable handbook is available through *Times Microwave*. Included are detailed specifications and full application notes on most cables.

Magic sinewaves are creating a revolutionary new power electronics development. They greatly improve the efficiency and can dramatically reduce the costs of such products as induction motor speed controls, home energy efficiency improvers, electric autos, and solar inverters. This is a *billion* dollar new opportunity that's largely up for grabs.

Quite a bit has happened on the magic sinewave front in the last few weeks. I'll be happy to send a free tutorial reprint to anyone who wants one. Formal proposals and a rather detailed tutorial library are also now available to serious inquirers only. These are also found on PSRT as files MAGSINT.PDF and MSPROP1. PDF.

Seminars, my PIC source code, and codeveloper programs are also newly available. Seminars are done in-plant or in a wilderness resort lodge. **EN**

*The NE602 could very well become the RF experimenter's "555" chip.
Learn about this fascinating and versatile device for your next RF project.*



JOSEPH J. CARR

Every now and then a chip comes along that strikes the public imagination, so it gets used in a lot of projects. The 741 operational amplifier was like that in the early 1970s. Also reaching a high pitch of popularity was the 555 IC timer chip. Both of those chips reached such heights because they were both useful and well-behaved (i.e. they did what they did with little muss or fuss). The radio frequency (RF) hobbyist, however, only recently found a chip that meets those requirements: the NE602 from Signetics.

The NE602 device is a monolithic integrated circuit containing a double-balanced mixer (DBM) and an internal oscillator circuit. The DBM has balanced inputs (pins 1 and 2), balanced outputs (pins 4 and 5), and can operate at up to 500 MHz. The internal oscillator circuit provides an emitter connection and a base connection to the outside world. Figure 1-a shows the block diagram, and Fig. 1-b the pinouts for the NE602 device.

The NE602 is meant to be used as the receiver front-end in VHF portable telephones, but a lot of amateur radio and electronics enthusiasts have used the chip for a wider variety of applications, some of which we'll talk about here. The NE602 is a strong candidate whenever you want to build a frequency converter or translator, or

even a signal-generator circuit. We can do that with oscillator circuits consisting of inductor-capacitor (L-C) variable-frequency oscillators, or piezoelectric crystals in either voltage-tuned or swept-frequency arrangements. We're going to explore some of the various configurations of circuits for the NE602 device, including the DC-power-supply connections, the RF-input configurations, the local-oscillator circuits, and the output circuits.

The NE602 version of the device operates over a temperature range of 0- to +70°C, while the related SA-602 device operates over an extended temperature range of -40- to +85°C. The most common form of the NE602, and most useful for the hobbyist and experimenter, is the NE602N, which is in an eight-pin mini-DIP package. An eight-lead surface-mount package (NE602D) is also available.

Heart of the NE602. Because the NE602 contains both a DBM and a local oscillator (LO), it can be used as the entire front-end of a radio receiver. Figure 2 shows a partial view of the internal circuit of the heart of the NE602: the double-balanced mixer stage. That configuration is known as a Gilbert transconductance cell. It consists of a pair of cross-coupled differential amplifiers. One feature of the design is that it offers a very good

USING THE NE602

noise figure, which is typically 5 dB at 45 MHz. The third-order intercept point is -15-dBm referenced to a matched input. Unfortunately, the dynamic range is not what it could be, so a good idea is to be sure that the input signal levels do not exceed -25 dBm (≈ 3.16 mW). That signal level is similar to about 12.6 mV into a 50-ohm load, or 68 mV into the 1,500-ohm input impedance of the NE602. The NE602 is capable of providing 0.2- μ V sensitivity without the need for external RF amplification. Although the straight NE602 suffers from dynamic range problems, the improved NE602A is said to solve that problem.

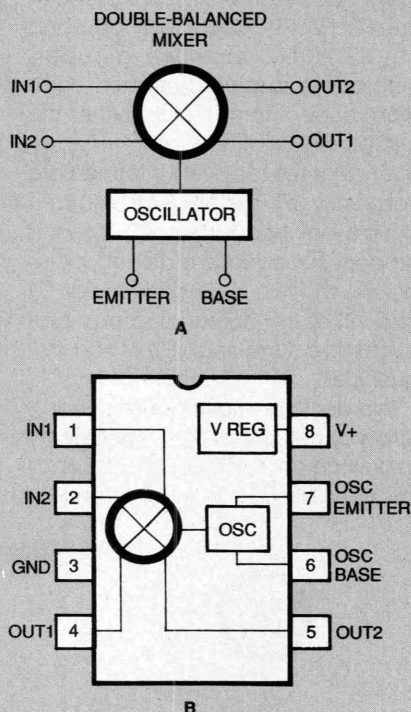


Fig. 1. The NE602 contains a double-balanced mixer and local oscillator. Here are its block diagram (A) and pinouts (B).

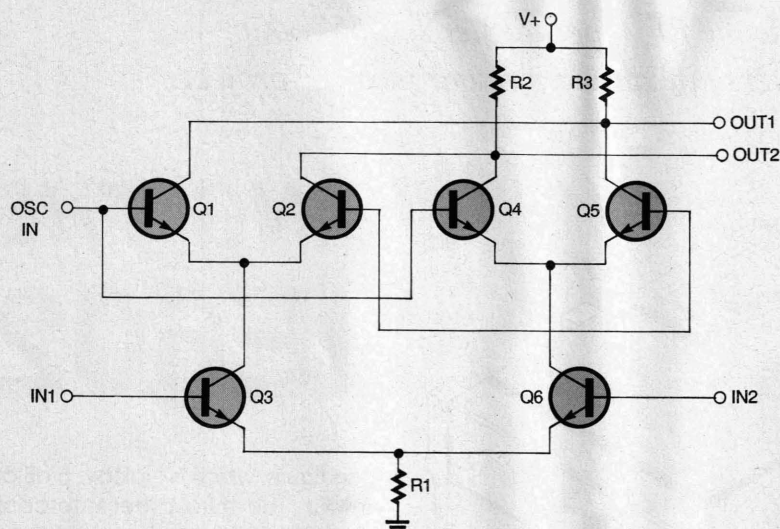


Fig. 2. The heart of the NE602: a Gilbert transconductance cell used for the double-balanced mixer.

Frequency Translation. The process of frequency translation or conversion is called heterodyning. When two frequencies (F_1 and F_2 in Fig. 3) are mixed together in a nonlinear circuit, a collection of different frequencies will appear at the output. Those frequencies are characterized as $mF_1 + nF_2$, where n and m are integers or zero (0, 1, 2, 3...). For the sake of simplicity, we normally consider only the cases where m and n are either 0 or 1, so the output frequencies are F_1 , F_2 , $F_1 - F_2$ (difference), and $F_1 + F_2$ (sum). To make a superheterodyne receiver (the most common modern form), select either the sum ($F_1 + F_2$) or difference ($F_1 - F_2$) frequency as the receiver's intermediate frequency (IF). The NE602 contains a double-balanced mixer, so when it is properly impedance matched, it suppresses the two input frequencies (F_1 and F_2) at the output, and only produces the sum and difference frequencies.

In order to provide frequency translation by heterodyning, it is necessary to provide an LO circuit. The LO circuit inside the NE602 consists of a transistor

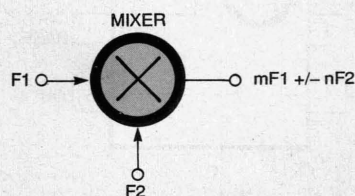


Fig. 3. In a basic mixer circuit, the sum ($F_1 + F_2$) and difference ($F_1 - F_2$) of two input frequencies appear at the output.

with its base and emitter elements available to the outside world. Oscillators using that circuit will operate up to 200 MHz. Any form of oscillator can be built, as long as the circuit does not need a connection to the collector of the oscillator transistor. Because of that restriction, both L-C and crystal variants of the Colpitts, Clapp, Hartley, Butler and other oscillator circuits can be built, while the Pierce and Miller circuits are not possible.

DC Power Supply Connections.

The power is applied to the NE602 between pins 3 (ground) and 8 ($V+$). The DC power supply voltage range is +4.5- to +8-volts DC, with a current drain ranging from 2.4 to 2.8 mA.

The DC power supply terminal (pin 8) must be decoupled with a 0.01- to 1- μ F capacitor (0.1 μ F is most common). The bypass capacitor must be mounted as close as possible to the body of the NE602, and must be capable of good performance at RF frequencies (some capacitors act like complex RLC networks at RF).

Figure 4 shows several possible DC power supply configurations for the NE602. In Fig. 4-a, the DC power supply voltage is between +4.5 and +8-volts DC, which is the normal operating range of the device. A resistor, usually 100 to 180 ohms, is placed in series with the $V+$ line to the NE602. If the circuit is operated from a 9-volt DC power supply (e.g. a 9-volt DC transistor-radio battery), then the resistor should be increased to a value between 1,000 and 1,500 ohms, as in Fig. 4-b.

If the DC power supply voltage is either unstable, or at a value higher than 9-volts DC, you might want to use some form of voltage regulation. In

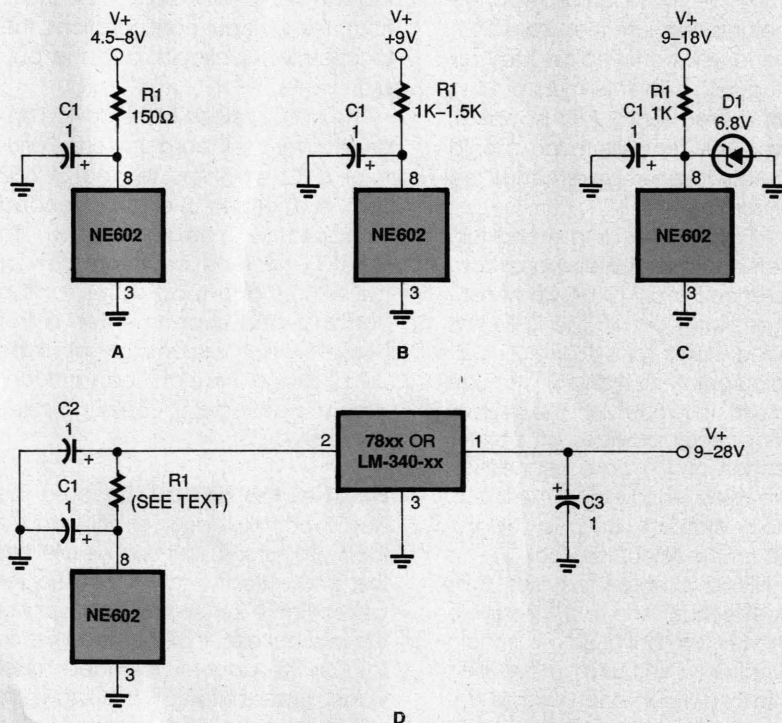


Fig. 4. There are several ways to power the NE602. A resistor should be placed in series between the power supply and the NE602 (a and b). A Zener diode (c) or a voltage regulator (d) can also be used.

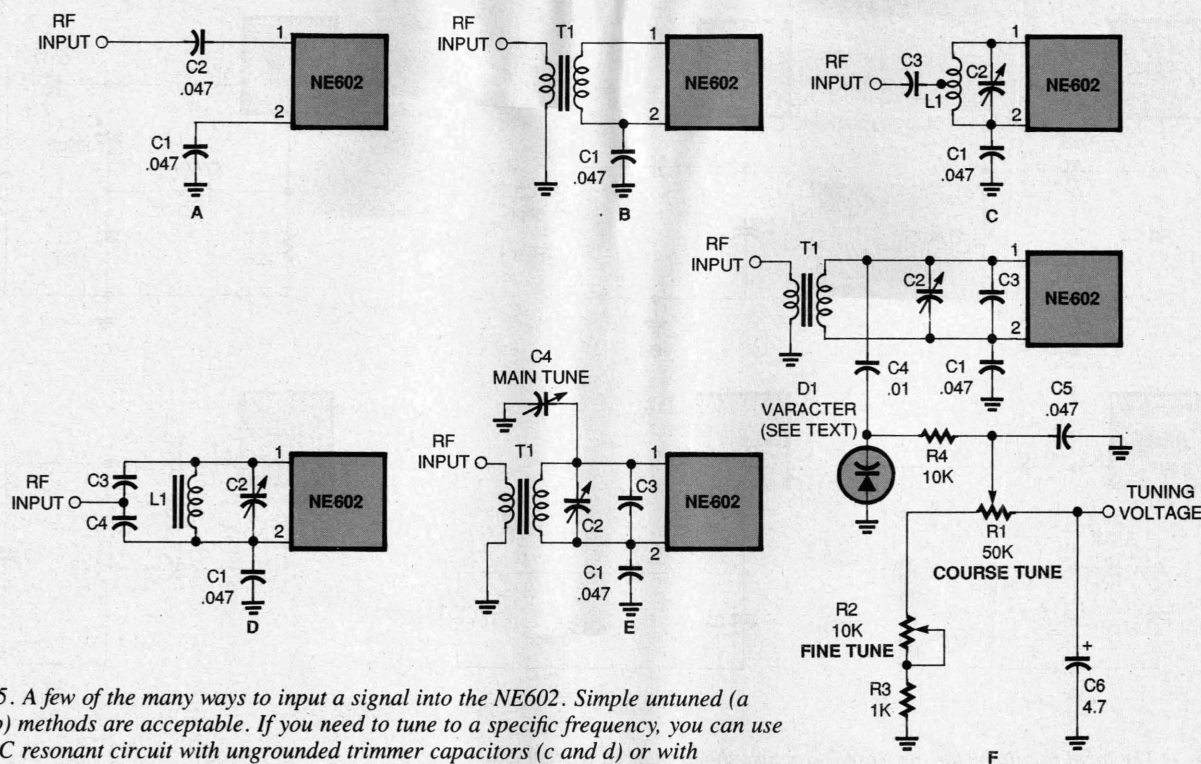


Fig. 5. A few of the many ways to input a signal into the NE602. Simple untuned (a and b) methods are acceptable. If you need to tune to a specific frequency, you can use an L-C resonant circuit with ungrounded trimmer capacitors (c and d) or with grounded variable capacitors (e). You can even use a tuning voltage in connection with a varactor (f).

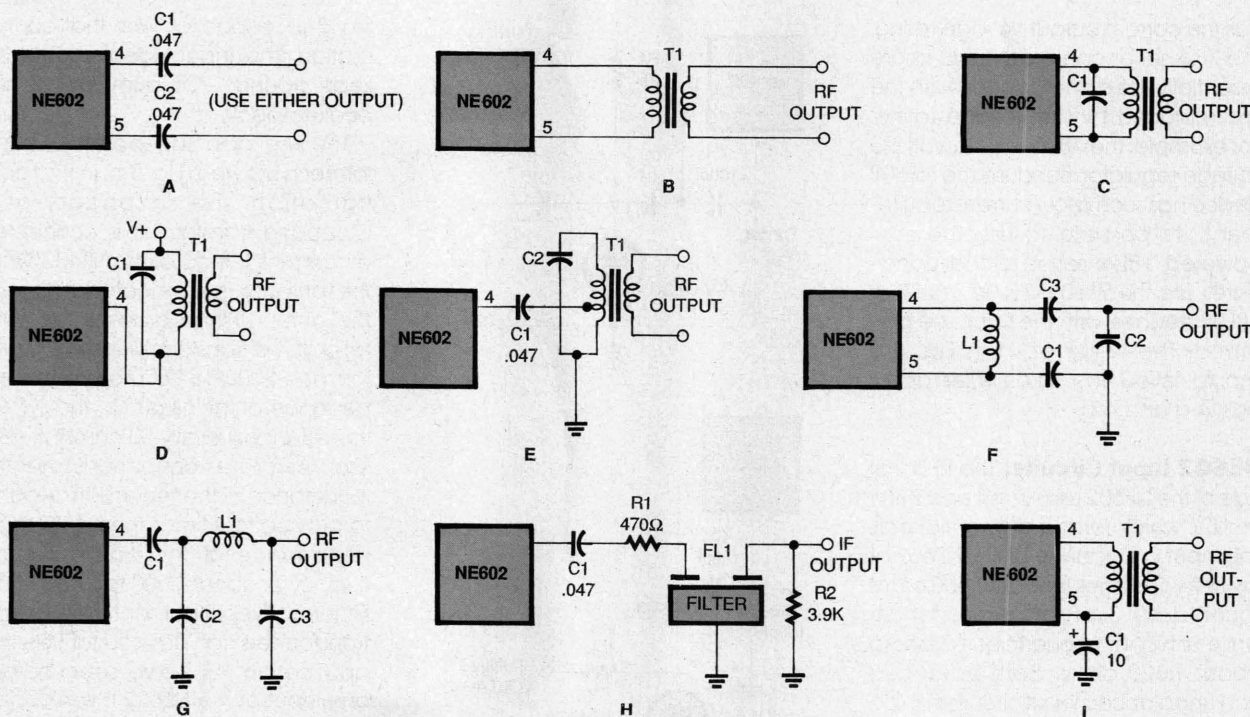


Fig. 6. The various output circuits shown here demonstrate how to either pass all the frequencies from the NE602, or allow only the sum or difference frequencies through, depending on which circuit is used.

fact, that's highly recommended. Figure 4-c shows the use of a Zener diode, rated at 6.8-volts DC, which keeps the supply voltage seen by the NE602 at that level even though the source power supply voltage might

vary from 9 to 18 volts, or so.

The use of a three-terminal voltage regulator is shown in Fig. 4-d. Those devices provide a constant output voltage for a wide range of DC input voltages. A typical voltage regulator

can accept input voltages from a minimum of about 2.5-volts higher than its rated output voltage, up to a maximum of about 30 to 38 volts. Almost any positive voltage regulator can be used in the circuit of Fig. 4-d if it

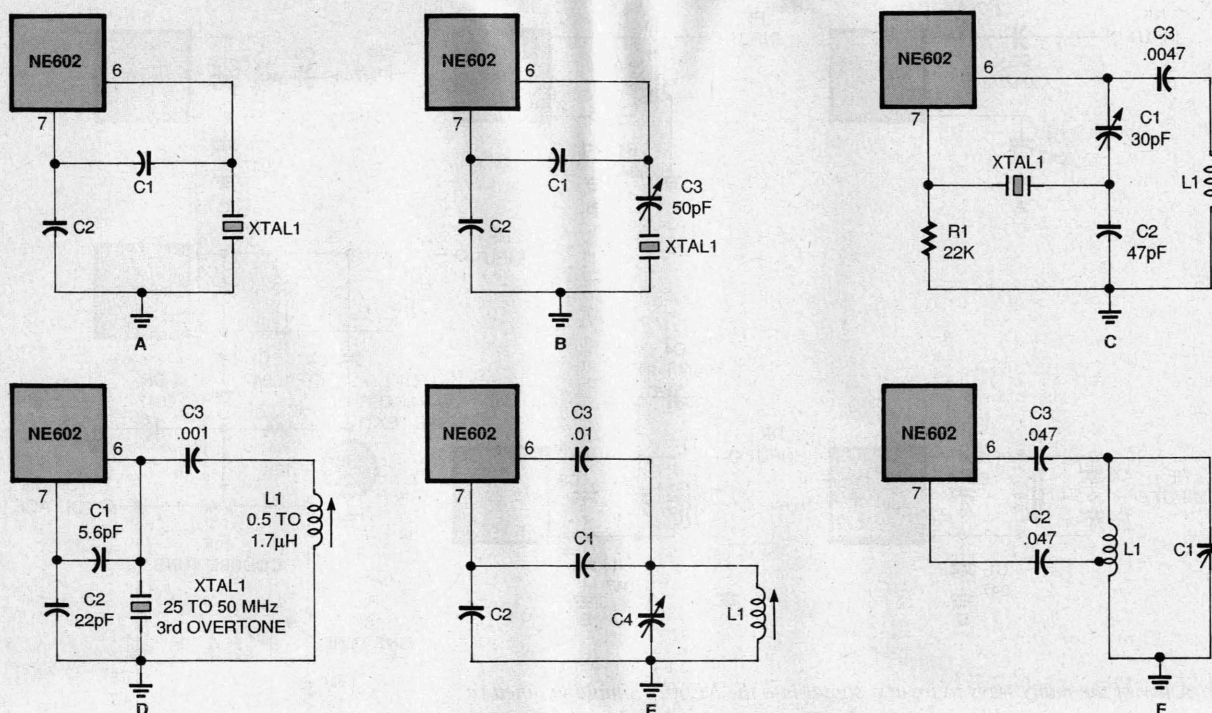


Fig. 7. The local oscillator for the NE602 can take the form of either a crystal-controlled (a through d) or a resonant-tank circuit (e and f).

has the correct output-voltage rating. The 78xx series and the LM-340-xx are essentially the same devices, with the "xx" replaced by the voltage rating. For example, the 7805 is a +5-volt DC voltage regulator. Because the NE602 device has such a low current requirement, it's possible to use the low-powered 78Lxx series. Good candidates are the 78L05, 78L06, 78L08 or 78L09 devices, with the first three preferred. The value of R1 in Fig. 4-d should follow the same rules as for Figs. 4-a and 4-b.

NE602 Input Circuits. The RF input side of the NE602 uses pins 1 and 2 (IN1 or IN2), which form a differential pair. The input impedance of the NE602 at lower frequencies is about 1,500 ohms shunted by 3 pF of capacitance, while at higher frequencies it drops to about 1,000 ohms. Both balanced and unbalanced input circuits can be used on the NE602. Unless a true differential configuration is used, the signal is usually applied to pin 1, and pin 2 is bypassed to ground.

Figure 5-a shows the simplest form of input circuit. A single capacitor (0.047- μ F typical) couples the signal from the outside world to pin 1 of the NE602. The other input pin (pin 2) is bypassed to ground with another

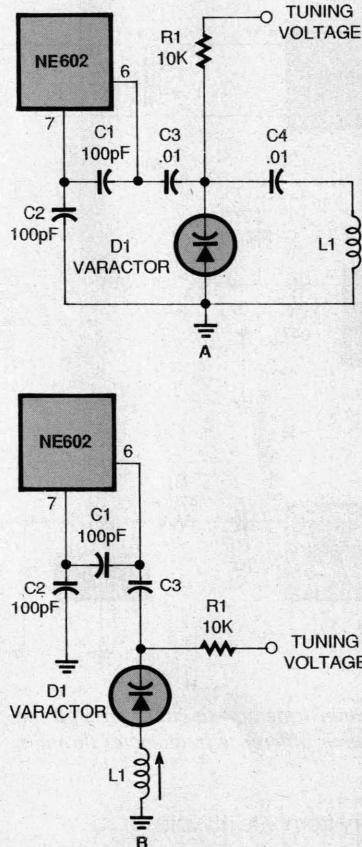


Fig. 8. The LO circuit for the NE602 can also be voltage-controlled. Here are two different methods of accomplishing that.

0.047- μ F capacitor. With that configuration, the input signal should be kept around -25 dBm, or 180-mV peak-to-peak.

The use of a wideband RF transformer is shown in Fig. 5-b. In that configuration, the secondary of a wideband transformer is connected across pins 1 and 2 of the NE602, while the primary is connected between the antenna and ground. The turns ratio of the transformer is set to transform the 1,000- to 1,500-ohm input impedance of the NE602 to the system impedance (usually 50 ohms). A general rule for the transformer is to set the inductance of the secondary winding to provide four times the NE602 input impedance at the operating frequency, or about 4,000 to 6,000 ohms. Either conventional- or toroid-wound transformers can be used for T1 in that application. As we've seen before, one input of the NE602 is decoupled to ground through a capacitor.

The circuits shown in Fig. 5-c and 5-d are tuned to a single frequency, but use different methods to provide impedance matching between the source and the input of the NE602. Of course, a capacitor could also be used to resonate the secondary of the transformer in Fig. 5-b. In Fig. 5-c, we do just that by resonating coil L1 with

variable capacitor C2. The input signal is coupled to coil L1 through an impedance-matching tap on L1 via capacitor C2. As with the other circuits, pin 2 is bypassed to ground.

The circuit in Fig. 5-d shows the use of a capacitor voltage divider (C3 and C4) to match the impedances. The resonant frequency is set by tuning L1 with variable capacitor C2, plus the series capacitance of C3 and C4. The tuning capacitance is:

$$C_{\text{tune}} = C2 + C3 \times C4 / C3 + C4$$

The circuits shown in Figs. 5-c and 5-d are used when the source impedance to be matched is less than the 1,500-ohm input impedance of the NE602 device. When a transformer input such as Figs. 5-b and 5-e are used, then the source impedance can be either higher or lower than 1,500 ohms, provided that we have the correct transformer turns ratio. One difficulty with the resonant circuits discussed earlier is the fact that the capacitor is connected across L1, which means that it must be ungrounded. That's fine as long as you can use trimmer capacitors, or somehow insulate a normally grounded variable capacitor. But it's not always practical to do that, so you can use the circuit of Fig. 5-e instead. Three tuning capacitors are used in that circuit: C2, C3 and C4. Capacitor C4 is effectively across the secondary winding of T1, provided that the value decoupling capacitor C1 is very large compared to the value of C4. In that instance, the value of the series combination of C1 and C4 is very nearly the value of C4. Capacitor C4 is used as the main tuning capacitor, while C2 is an optional trimmer capacitor for fine tuning. Also optional is C3, which is used to make up any extra capacitance required to meet a minimum value. If $C1 \gg C4$, the effective capacitance across the secondary of T1 is $C2 + C3 + C4 + 3 \text{ pF}$.

A voltage-tuned variant of the input

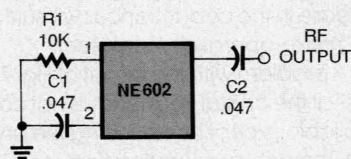


Fig. 9. If LO signal of the NE602 is sent directly to the output pins, the device can be used as a low-cost, high-frequency oscillator.

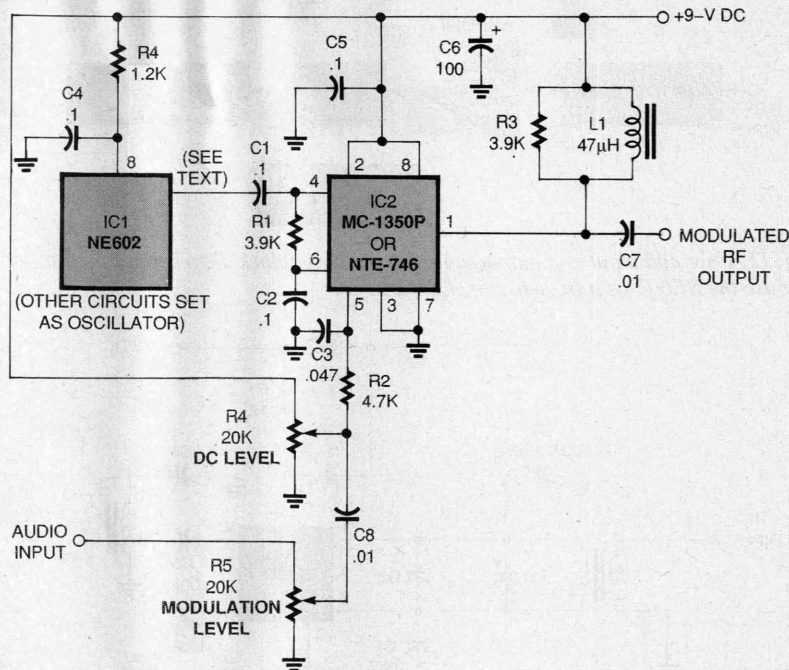


Fig. 10. By using an MC-1350P modulator IC, the output of the NE602 can easily be amplitude modulated.

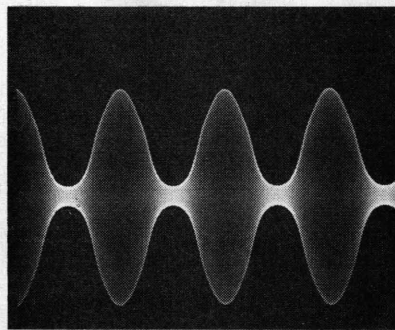


Fig. 11. This is the waveform of the circuit in Fig. 10, ready to be transmitted as an AM-radio signal.

circuit is shown in Fig. 5-f. That circuit is similar to Fig. 5-e, except that the variable "main tuning" capacitor (C4 in Fig. 5-e) is replaced with a voltage variable-capacitance (varactor) diode, D1. Those diodes provide a capacitance that drops as you increase the reverse voltage across the diode's terminals. As long as C4 is very much larger than the capacitance of D1 at any voltage, the tuning capacitance is that of D1. Resistor R1 is used to isolate the tuning voltage from the diode, so that it doesn't load the capacitance. Although a single "main tuning" potentiometer could be used, the circuit uses both coarse tune (R1) and fine tune (R2) potentiometers.

NE602 Output Circuits. The NE602

has two outputs that can be used as a balanced pair, or alone as single-ended outputs. Either pin 4 or pin 5 can be used alone for single-ended circuits, or pins 4 and 5 are used together as a balanced, or differential, output. The simplest form of output circuit is shown in Fig. 6-a. Either pin 4 or pin 5 can be used. The output signal is coupled through a 0.01- to 0.1-µF capacitor to whatever circuit the NE602 is to drive. The circuit of Fig. 6-a is a wideband configuration and cannot tell which signals are the sum ($F1 + F2$) or difference ($F1 - F2$) signals.

A wideband balanced-output circuit is shown in Fig. 6-b. The transformer is used to change the 1,500-ohm output impedance of the NE602 to the impedance of whatever circuit is being driven. If both input and output system impedances are the same, then the same type of transformer can be used for both input and output, although reversed relative to each other. As in the case of the input circuit, either standard or toroidal transformers can be used for T1.

Tuned-output circuits are shown in Figs. 6-c through 6-e. The circuit of Fig. 6-c is balanced. The primary of the transformer is connected across pins 4 and 5, and is resonated by capacitor C1. A single-ended variation is shown in Fig. 6-d. In that case, the parallel-tuned circuit consists of C1

that can be specified if crystals are being ordered directly from a manufacturer. In Fig. 7-b, a variable, or "trimmer" capacitor is placed in series with the crystal in order to set the frequency. The trimmer capacitor can be adjusted to set the oscillator to the desired frequency.

The two previous crystal oscillators operate in the fundamental mode of crystal oscillation. The resonant frequency in the fundamental mode is set by the dimensions of the slab of quartz used for the crystal; the thinner the slab, the higher the frequency. Fundamental-mode crystals work reliably up to about 20 MHz, but at higher frequencies the slabs become too thin for safe operation; at that point, the thinness of the slabs of fundamental-mode crystal causes them to fracture easily. An alternative is to use overtone-mode crystals. The overtone frequency of a crystal is not necessarily an exact harmonic of the fundamental mode, but is close to it. The overtones tend to be close to odd

integer multiples of the fundamental (3rd, 5th, 7th, etc.). Overtone crystals are marked with the appropriate overtone frequency, rather than the fundamental.

Figures 7-c and 7-d are overtone-mode crystal-oscillator circuits. The circuit in Fig. 7-c is a Butler oscillator. The overtone crystal is connected between the oscillator emitter of the NE602 (pin 7) and a capacitive voltage divider that is connected between the oscillator base (pin 6) and ground. There is also an inductor in the circuit (L1) that must resonate with C1 to the overtone frequency of crystal XTAL1. Figure 7-c can use either 3rd- or 5th-overtone crystals up to about 80 MHz. The circuit in Fig. 7-d is a third-overtone crystal oscillator that works from 25 to about 50 MHz, and is simpler than Fig. 7-c.

A pair of variable-frequency oscillator (VFO) circuits are shown in Fig. 7-e and 7-f. The circuit in Fig. 7-e is a Colpitts-oscillator version, while Fig. 7-f is a Hartley-oscillator version. In both

oscillators, the resonating element is an inductor-capacitor (LC) tuned-resonant circuit. In Fig. 7-e, however, the feedback network is a tapped-capacitor voltage divider, while in Fig. 7-f it is a tap on the resonating inductor. In both cases, a DC-blocking capacitor to pin 6 is needed in order to prevent the oscillator from being DC grounded through the resistance of the inductor.

Voltage Tuned NE602 Oscillator Circuits.

Figure 8-a and 8-b show a pair of VFO circuits in which the capacitor element of the tuned circuit is a voltage-variable capacitance diode, or varactor (D1 in Fig. 8-a and 8-b). Those diodes exhibit a junction capacitance that varies in direct response to the reverse-bias voltage applied across the diode. Thus, the oscillating frequency of those circuits is controlled by a tuning voltage. The version shown in Fig. 8-a is a parallel-resonant Colpitts oscillator, while Fig. 8-b is a series-tuned Clapp oscillator.

Using the NE602 as a Signal Generator.

The NE602 is normally used as a receiver front-end or as a frequency converter. It can also be used as a signal generator. Figure 9 shows the basic configuration for providing the LO signal at output pins 4 and 5; place a 10,000-ohm resistor (R1) between pin 1 and ground, while bypassing pin 2 to ground through a 0.047- μ F capacitor. The output signal is taken from either pin 4 or 5 through another 0.047- μ F capacitor.

The output signal of Fig. 9 will be a sine wave at the frequency of oscillation for the oscillator circuit connected to pins 6 and 7. That signal can be swept or frequency modulated by using one of the varactor LO circuits shown in Fig. 8. To sweep the frequency, make the tuning voltage a sawtooth waveform, while to frequency-modulate it, use a sine wave. If you want to amplitude-modulate the signal, then use a circuit such as Fig. 10.

The signal source is any of the NE602 oscillators (IC1 in Fig. 10), while the modulator is IC2, an MC-1350P chip. That chip is also available from the service-repair industry replacement lines as the NTE-746 or ECG-746. It is an RF-gain block with a gain-control terminal (pin 5), and that gain-control terminal can be used for the

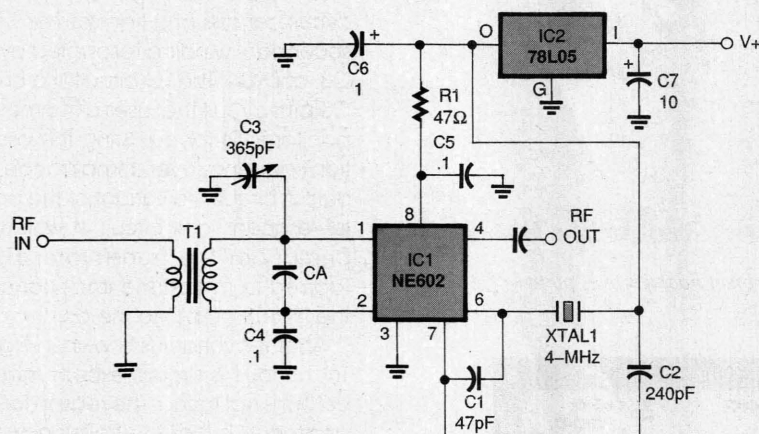


Fig. 14. Here is another simple frequency translator circuit; it does not select which frequency appears at the output.

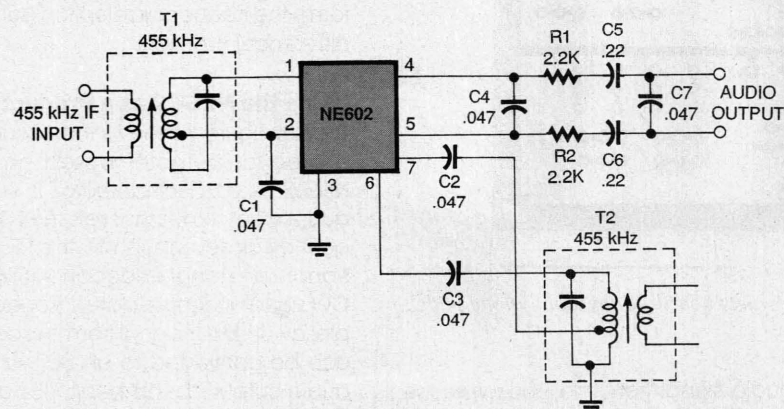


Fig. 15. After you've tuned in a particular radio frequency and demodulated it, this product-detector circuit can be used for Morse code (CW) or single-side band (SSB) reception.

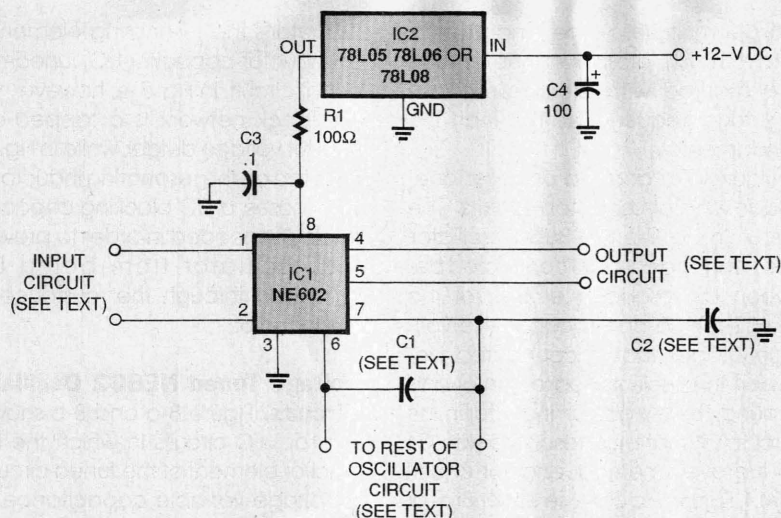


Fig. 16. There is just enough circuitry on the universal project board for NE602 to get you up and running with many different projects.

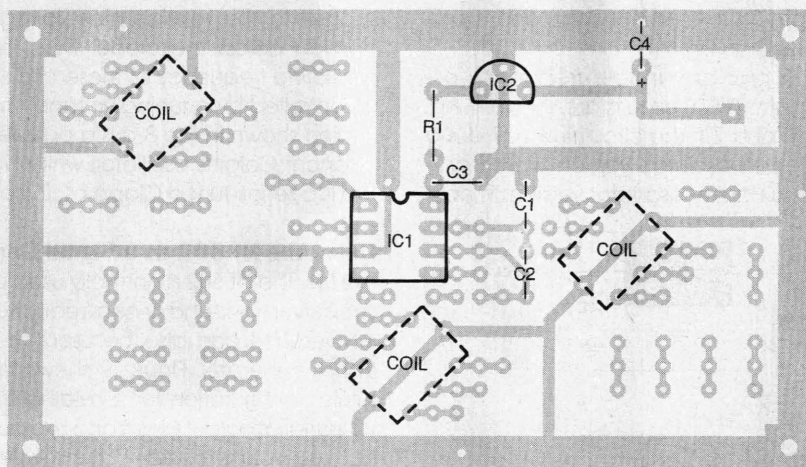
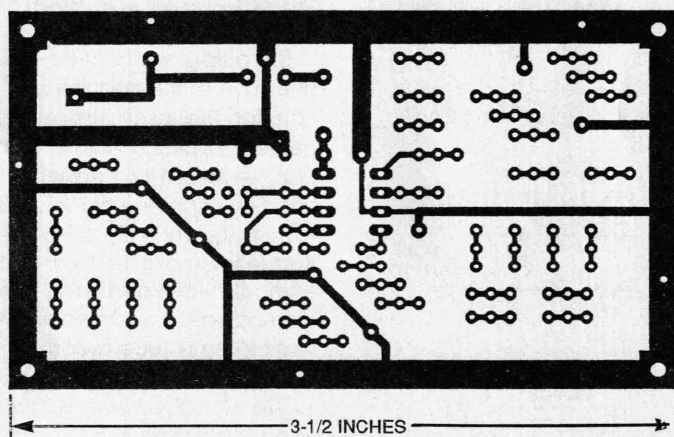


Fig. 17. Component placement on the universal project board leaves plenty of room for your own circuitry.



Here's the foil pattern for the NE602 universal project board. Many stand-alone pads are included to hold your circuitry.

amplitude-modulation function. Two signals are applied to the gain-control terminal as shown in Fig. 10; a DC level from potentiometer R4, and the

audio signal from the MODULATION LEVEL control (R5). Adjust both DC LEVEL and MODULATION LEVEL until the output signal, as viewed on an oscilloscope, looks

like Fig. 11. There should be good symmetry and no clipping of the peaks.

Using the NE602 in Converter Projects.

The basic frequency converter was shown in block form back in Fig. 1-a; now it is time to expand upon that as shown in Fig. 12. The NE602 and its supporting circuitry might be used as the mixer and local oscillator. The amplifier is optional, and should only be used when sensitivity is poor. Poor sensitivity might be caused when insertion loss of the input tuning network is high.

The input and output tuning networks are used to segregate the signals. The input tuning selects the desired RF-input frequency (F1), and rejects all other frequencies. The output tuning network selects either F1 - F2 (difference), or F1 + F2 (sum) signals. Those filters can be L-C resonant-tank circuits, low-pass filters, high-pass filters, or band-pass filters as needed for the specific application.

Figure 13 shows a basic NE602 frequency-converter circuit. The input circuit consists of a transformer with a secondary winding resonated by C2, C3, and C4. The LO circuit is a crystal Colpitts circuit that uses a trimmer capacitor (C9) for adjusting the oscillation frequency over a small range. The output circuit is a variant of the parallel-resonant tank circuit in which the primary of the transformer (T2) is tapped to match the impedance of the NE602 output to the coil.

Another variant is shown in Fig. 14. That circuit is similar, except that the output is not tuned. The reason for that approach is that the frequency converter is used to drive the antenna input of a radio receiver, which performs the frequency selection (sum vs. difference) function.

Using the NE602 as a Product Detector.

Figure 15 shows the circuit for a product detector based on the NE602. A product detector is a frequency converter that sets the LO frequency close enough to the RF or IF signal so a single-sideband (SSB) or CW signal is demodulated. For example, a 455-kHz IF signal from a receiver can be converted to an 800-Hz CW audio output ("beep-beep") by using an LO at 455.8 kHz or 454.2 kHz. The difference tone is found in the output.

(Continued on page 79)

BUILD THE PCDRILL

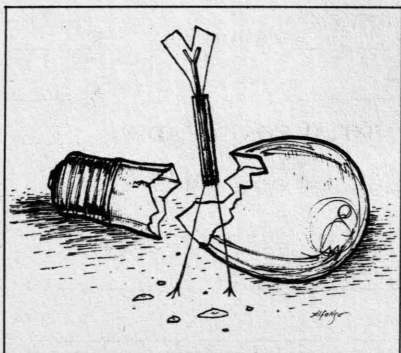
(Continued from page 46)

the compiled version. (Look for PCDRILL.ZIP on the Gernsback FTP site —ftp.gernsback.com/pub/EN).

Select each of the first four options in turn. The table and the drill should move in the selected direction. The test program moves the table and caddy in "slow motion," but real applications (like the one that will be presented next time) produce quicker and smoother movement. If either motor stalls and the table or drill does not move, the associated guide is probably jamming against the track. Adjust the guide and try again. Repeat the process until both table and caddy move smoothly throughout the entire range of travel.

Several tweaks can improve the performance of PCDrill. For instance, apply paste wax to all moving wooden surfaces, including the drill caddy sides and drill slides. All moving metal surfaces should get a thin coat of light oil. That includes the driver nuts, threaded rods, and aluminum angle tracks. Those lubrication efforts will decrease friction and make for smoother operation.

Next time we'll align PCDrill, and introduce the application program and all of its functions. We'll also look at the application's data file and explain each of its entries, including the Speed entry that maximizes speed of movement. We'll also provide details for building a 5.75-volt power supply from scratch, and use PCDrill to fabricate its own PC board. To round things out, we'll provide an AC wiring option that allows you to energize the drill only during actual drilling, as opposed to having the drill run continuously. See you then. Ω



CONDUCTANCE ADAPTER

(Continued from page 63)

keep the input leads short to avoid 60-Hz line pickup. Although the ICs used in the Conductance Adapter are ESD protected, you should avoid letting any static discharges into J1 or J2.

Connect a 1-megohm resistor to J1 and J2 and the output on a DVM will be 1.000 volts, which is equal to 1.00 micromho or 1000 nanomhos. A 300-megohm resistor will read 3.33 millivolts, which is equal to 3.33 nanomhos. A diode (shown as D_x in Fig. 1) might show a reading of 2.55 millivolts when reverse biased. That is equal to a leakage current of 2.55 nanoamps. Leakages can be measured down to 10 picoamps. If you need a voltage greater than -1.00 volt for your tests, do not make any connection to J1. Apply an external voltage between the ground jack (J3) and the free lead of the device being tested. Do not let any voltages applied to J2 exceed 3 volts, or IC2 will be destroyed. Ω



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The Positive Place For Kids

USING THE NE602

(Continued from page 54)

If the signal is an SSB signal, then the LO is set at a frequency of 2.5- to 2.8-kHz higher or lower than the IF, depending on whether you want to demodulate an upper-sideband (USB) or lower-sideband (LSB) signal.

The input signal circuit in Fig. 15 uses a 455-kHz IF transformer of the sort used for transistor radios (see Digi-Key or Mouser catalogs for suitable types). The transformer that you want to use for T1 is the type that has a resonant secondary with a tapped inductance. The LO circuit uses the same type of transformer as the input, but configured as a Hartley oscillator.

The output signal is at audio frequencies, and is filtered by an R-C network. The audio output is balanced, so it should be fed to a differential audio amplifier such as an op-amp.

NE602 "Universal" Project Board.

We've included a printed circuit pattern for a "universal" project board based on the NE602. The on-board circuitry (Fig. 16) is limited to the DC power connection, which is regulated by a three-terminal IC voltage regulator. All other functions can be set by you to make any project that you want. There are a large number of multi-pad stand-alone connections for various components depending on the circuit that you want to make, as well as positions for three six-pin standard shielded coils of the sort manufactured by Toko and sold by Digi-Key. You can also use these same holes for mounting home-brew toroid inductors. Figure 17 shows the parts placement of the universal project board. The universal NE602 board can be bought from FAR Circuits, 18N640 Field Court, Dundee, IL, 60118 for \$4 plus \$1.50 shipping for every four boards ordered (i.e. 1 to 4 boards shipped for \$1.50). IL residents will have to add appropriate sales tax.

Now you've seen how well-behaved the NE602 is. Here is an RF chip that will function in a variety of applications from receivers, to converters, to oscillators, to signal generators. With the universal project board, the task of testing a new design based on the NE602 becomes "duck soup." Ω

Analog Telemetry Techniques

— while designed for biomedical signals, these circuits work with any analog data

WARNING: Use or sale of this or similar devices is restricted under Federal Law to physicians or on their orders. No attempt should be made to diagnose or treat patients without medical supervision.

Joseph J. Carr K4IPV
5440 South 8th Road
Arlington VA 22204

The use of OSCAR to transmit human electrocardiograms (ECG—or EKG, after the German spelling), and the article in the February, 1978, issue

of 73 Magazine ("Inexpensive EKG Encoder" by WA3AJR), indicate interest in the transmission of analog data. The techniques needed to transmit the human ECG are also useful for transmitting almost any form of analog data signal in which the repetition rate is low and all frequency components

fall into the "under 500 Hz" range. In this article, we will examine such circuits from the viewpoint of the human ECG and certain other physiological signals, but the information can be applied to other problems as well.

The ECG Preamplifier

The peak amplitudes in the ECG waveform are on the order of 1 or 2 millivolts, and significant frequency components exist in the 0.05- to 100-Hertz range. Electronic amplifiers used in professional diagnostic ECG equipment will have this AHA-recommended frequency response, while equipment used exclusively for monitoring usually has a frequency response of 0.05 to 40 or 50 Hertz (depending upon manufacturer).

ECG preamplifiers usually are differential types so that their inherent rejection of common mode signals can be used to elimi-

nate 60-Hertz interference.

Additionally, ECG preamps must be capacitor-coupled to prevent drift due to changes in the *electrode-offset potential* that exists between the electrode and the patient's skin. The electrolytic skin surface reacts with the metallic electrode to form a battery that produces upwards of a volt or more.

Fig. 1 shows the circuit for an ECG preamplifier used in a battery-operated radio telemetry transmitter. This circuit operates from a single 9- to 12-volt dc power source, although, in most non-portable applications, it could easily be redesigned to take advantage of the operational amplifier's bipolar power supply terminals.

The circuit in Fig. 1 is basically an ac-coupled version of the classic instrumentation amplifier, and has a voltage gain of approximately 1000 (see "Op-Amp Encyclopedia"

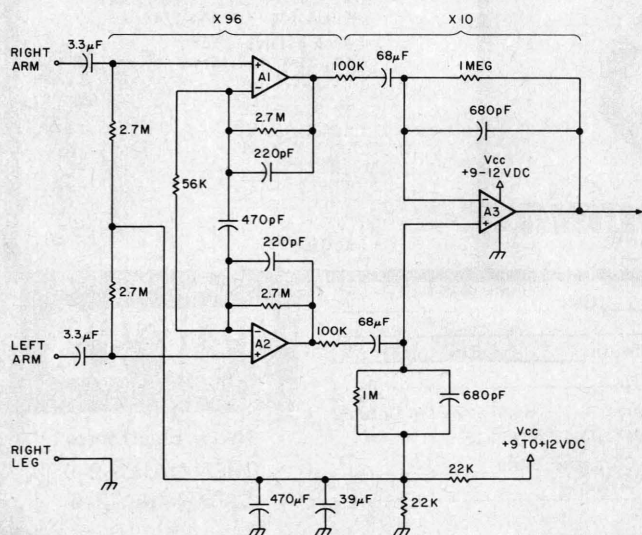


Fig. 1. Typical ECG preamplifier.

by K4PV, 73 Magazine, February, 1978).

Under specific sets of circumstances, ac mains-powered equipment can be hazardous to connect to some patients, so modern ECG preamplifiers use isolated designs. In most cases, the output of a preamplifier such as in Fig. 1 will be used to amplitude-modulate an oscillator operating in the 30- to 100-kHz range (see Fig. 2). The preamplifier will be on an isolated portion of the printed circuit board, along with the amplitude-modulator and a floating rectifier-filter that creates a dc power supply from the 50-kHz signal. Alternatively, some use separate dc-to-dc converters that do essentially the same thing from a separate oscillator.

The situations that create a hazardous environment for patients do not ordinarily exist outside of the hospital/medical environment, but experimenters and science-fairists who might want to use these techniques are advised *not* to connect people to any mains-powered equipment. The use of battery power provides the same isolation as used in professional equipment, but only if the entire apparatus is battery operated.

Transmission Encoders

Some portable units do not encode the ECG waveform at all before transmission, and are called direct-FM telemetry systems. In this type of transmitter, the output of a preamplifier such as in Fig. 1 is applied directly to a varactor-modulated crystal oscillator. In one popular brand, the oscillator operates at 12 to 14 MHz and is multiplied into the 174- to 215-MHz region with the final deviation being 75 kHz or so.

Other models first en-

code the ECG waveform in the form of a frequency-modulated audio carrier that can be transmitted over a wire or radio communications channel. This method is called FM/FM telemetry. The encoder consists of a voltage-to-frequency converter, or voltage-controlled oscillator (vco), that uses the ECG signal as its control or input voltage. The output of the vco is then applied to an ordinary FM transmitter.

Fig. 3(a) shows a circuit for an ECG encoder that was originally designed at the National Institutes of Health (NIH)¹ in Bethesda MD, for use in telephone call-in cardiac pacemaker-surveillance systems.² The circuit was built into the case of a modem (modulator/demodulator used in computer systems), and was issued to patients. The box had a pair of telephone earpiece cups so that the tone could be transmitted down the line. The patient would phone in to the pacemaker clinic once a week, where an analog recording of the waveform was made, and certain computer measurements were taken. These were

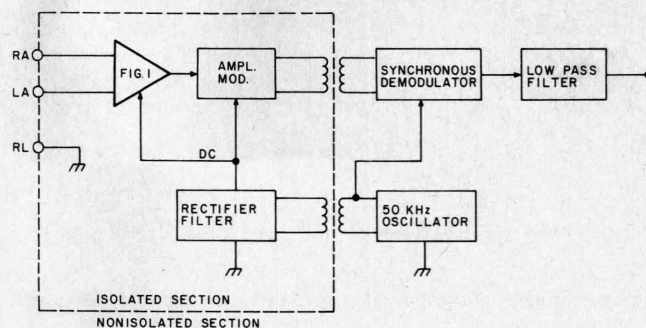


Fig. 2. Block diagram of an isolated ECG preamplifier.

reviewed later by a cardiologist (an M.D.), who would decide whether or not to ask the patient to come in for a closer examination.

The encoder in Fig. 3(a) uses four 741-family operational amplifiers and a Motorola MC4024P vco chip that drives a loudspeaker in the modem box. Note that this device is filtered too heavily to provide diagnostic quality ECG recordings, but is sufficient for the limited purpose intended. For those who wish to use the circuit for a wider frequency-response system, some modification of the low- and high-pass filter sections is in order.

Gain is provided by amplifier A1, while A2 serves as the high- and low-pass filter section. The fre-

quency response was limited intentionally so that interfering signals from the patient's skeletal muscles did not obscure the ECG waveform—a necessity because of the less-than-optimum situations in which recordings were attempted. The amplifier input lines were connected to a pair of 1.5-inch metal disk electrodes on the top surface of the modem box. This is not the best possible configuration, but is convenient for the home patient.

The 60-Hertz power mains will cause a tremendous interference signal to be present in nondifferential amplifiers and in those differential amplifiers in which the input balance cannot be maintained. To overcome this, a 60-Hz notch filter, A3 plus the cir-

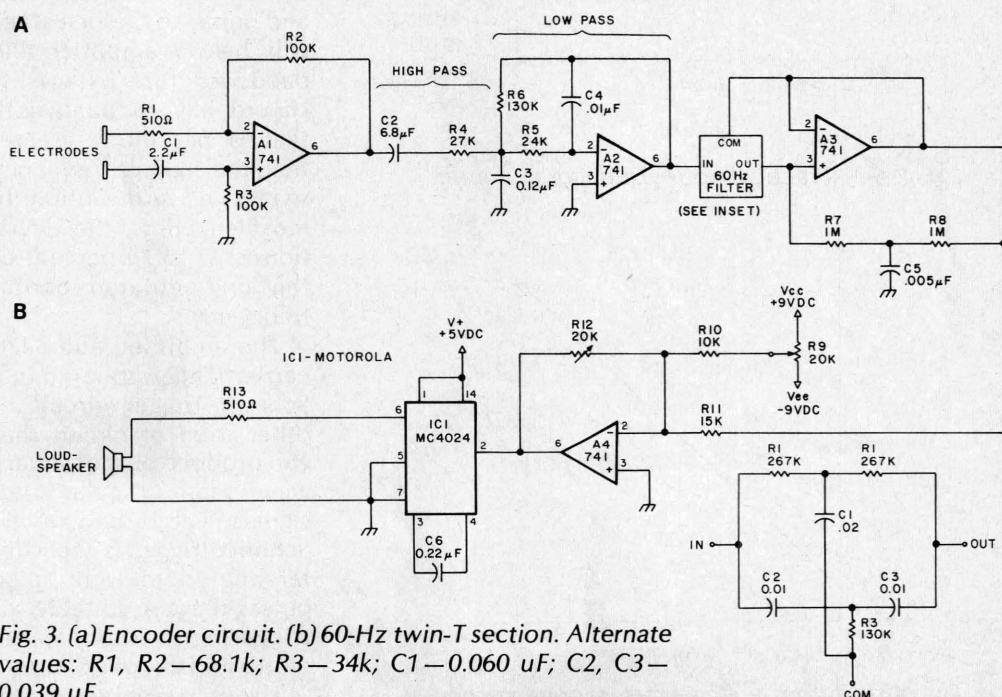


Fig. 3. (a) Encoder circuit. (b) 60-Hz twin-T section. Alternate values: R1, R2—68.1k; R3—34k; C1—0.060 uF; C2, C3—0.039 uF.

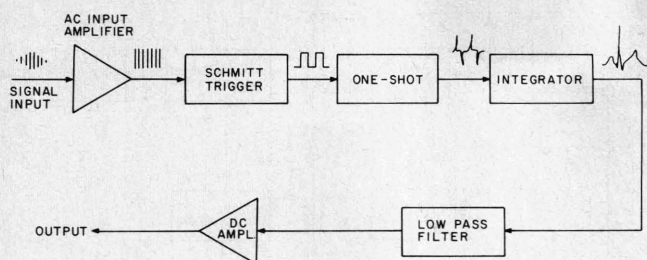


Fig. 4. Block diagram of a pulse-counting FM detector.

cuit in Fig. 3(b) is provided. A claim of 50-dB suppression is made, but we could verify only 18-23 dB.

Amplifier A4 serves as a level shifter between the ECG preamplifier and the MC4024P vco input. The output of the vco is fed to a loudspeaker for acoustical coupling to a telephone. Note that a series resistor is used between the speaker and the vco. This component is needed because the TTL output of the vco will be loaded too much by the low impedance typical of speakers.

The output of the vco in Fig. 3(a) is too high for most

radio transmitter audio input stages, and the waveform consists of a chain of square waves. To overcome this problem, an attenuator/low-pass filter is needed between the vco output and the radio transmitter input connector.

Decoders

Neither the circuit of Fig. 3(a) nor the circuit presented in the February, 1978, issue of 73 is useful by itself, except to show that your heart can control the whistle of the modulation in the loudspeaker! Unless a decoder (i.e., FM demodulator) is provided,

all these circuits can do is whistle at you in step with the ECG waveform!

The encoded signal is an audio-range (1 to 5 kHz) carrier that is frequency-modulated by the ECG waveform. Any phase-sensitive detection scheme could conceivably work, but two types are most commonly employed: phase-locked loops, and pulse-counting (also called digital) FM demodulators.

A number of technical papers have been written that show the use of low-cost PLL chips for ECG telemetry decoding, but, in my experience, these circuits have not been altogether satisfactory. It seems that many of the low-cost PLL chips do not recover rapidly enough following the ECG R-wave (the spike-like feature). This results in distortion of the waveform in an area that is of particular interest to the physician. Such circuits should, however, prove interesting to the hobbyist who can tolerate some distortion.

It is a little easier to use the pulse-counting detector of Fig. 4. The signal received from the radio loudspeaker or telephone line will probably be weak and noisy, so the first stage will be an amplifier and bandpass filter stage. It should have a bandwidth that is not much greater than the signal's frequency swing, i.e., 2x deviation. In most encoders, the deviation is 25 to 75 percent of the unmodulated carrier frequency.

The amplified audio-FM carrier is then squared in a Schmitt trigger circuit or other form of circuit that will produce output square waves from irregular input signals. The output of the Schmitt trigger is then differentiated to form spike pulses that are suitable for triggering a monostable multivibrator (one-shot).

The actual detector con-

sists of the one-shot and an integrator stage. This arrangement is common not only in telemetry applications, but also in many industrial and scientific instruments, including FM-carrier tape recorders that record low frequency analog data on ordinary audio tape. This circuit produces a dc level that is proportional to the frequency or pulse-repetition rate of the input signal.

The one-shot produces an output pulse for each trigger pulse received at its input. The one-shot output pulses differ from the input pulses, however, in that they have a constant amplitude and duration (period). Only their repetition rate varies with the input frequency. As a result, the output of the integrator, which is a time-averaging circuit, is a dc level proportional only to frequency. Note that at least one high-quality hi-fi FM tuner uses this technique at 10.7 MHz to demodulate the FM i-f signal. The integrator is a form of low-pass filter, so the circuit is inherently low-noise.

A practical example of this type of detector is shown in Fig. 5(a). Note that this is not a clinically acceptable circuit, but it is able to produce results that are good enough for educational or experimental applications. Most serious experimenters wishing to transmit low-frequency human or animal physiological signals by radio or wire, or to store them on an audio tape, will be successful with this circuit.

An LM311 voltage comparator is the input signal conditioner. This circuit is connected as a zero-crossing detector and requires at least 200 mV of signal to operate reliably. Lower signals can be accommodated if IC1 is preceded with an operational amplifier gain stage.

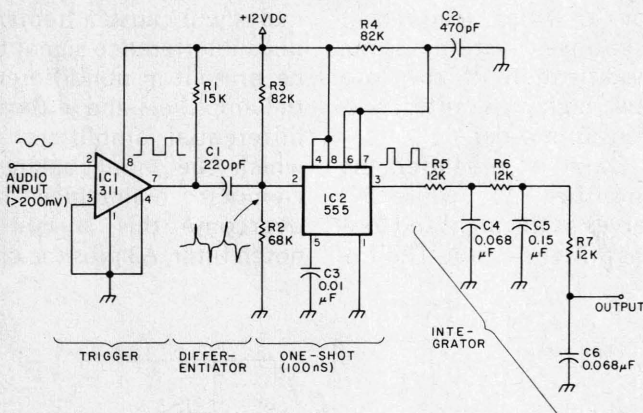


Fig. 5(a). Practical hobbyist decoder circuit.

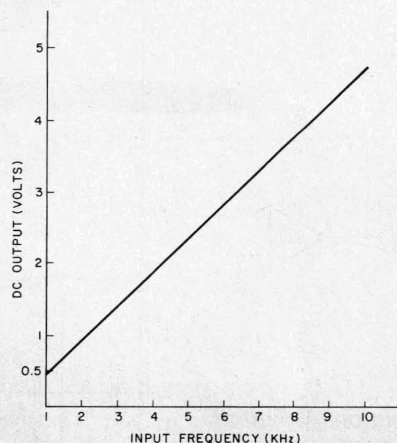


Fig. 5(b). Output voltage versus input frequency.

The square waves at the output of the LM311 are differentiated by C1-R2 to form spike pulses that will trigger the 555 one-shot stage (IC2). The 555 output pulses have a constant amplitude and duration, but their repetition rate varies with the input frequency.

The 555 pulses are integrated in a three-stage RC integrator. The output is taken across capacitor C6, and is a frequency-dependent dc potential. The graph showing input frequency versus dc voltage is shown in Fig. 5(b).

Does it work? I built a modulator using the A4/IC1 portion of Fig. 3(a) and used a function generator to drive its input. A triangle waveform of 2 Hz was selected because it approximates the frequency components of the ECG waveform. The output of the modulator was attenuated and then applied to the input of the circuit in Fig. 5(a). Fig. 5(c) shows the original input waveform from the function generator (upper trace) and the dc output of the decoder (lower trace). Note that the demodulated version shows some, but not much, loss of high frequencies in the waveform.

The vertical gain of the oscilloscope used was adjusted to show more clearly the two waveforms. The amplitude of the upper trace was several volts p-p, while that of the lower was approximately 80 mV p-p. An operational amplifier following the integrator would build this level up to whatever level is required.

Accessories

A qrs-beeper can be used by using the R-wave to trigger an audio oscillator. This is the "beep-beep" used to good effect in doctor TV shows. A popular method is to use the R-wave to fire a one-shot that drives one input of a NAND gate. The other input of the NAND gate is driven by a 1-kHz (or so) audio oscillator.

Alternatively, the output of the one-shot that is triggered by the R-wave can be connected to an LED to form a qrs-flasher. These circuits are shown in Fig. 6. In both cases, the noninverting input of the comparator might either be grounded, as in Fig. 5(a), or connected to a dc level that prevents the comparator from firing on noise impulses or, in most cases, the ECG T-wave.

The circuit of Fig. 5(a) is also useful to make a cardiometer (jargon for heart-rate meter) if the time constants are changed. The ECG waveform has a fundamental frequency of approximately 0.5 to 2 Hz, so it will not work properly with the 2-kHz values used when the decoder in Fig. 5(a) is used, but the same principle is used in the tachometer. (The same principle has been published many times as an auto tachometer, incidentally, again with suitable component value changes).

In the case of a cardiometer, the one-shot duration should be 25 to 50 ms, and the values of the integrator capacitors and resistors will probably have to be raised. The dc output meter can be

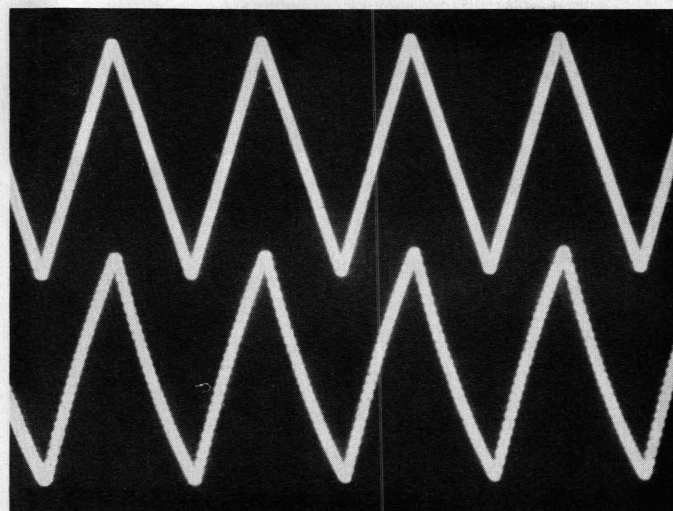


Fig. 5(c). Input (upper) and output (lower) waveforms.

calibrated in beats-per-minute. If you do not have a 1-Hz range function generator to calibrate this circuit, then use TTL or CMOS counter/dividers or a crystal calibrator circuits to divide the 60-Hz line from a 5-V ac filament transformer (not direct from the 115 V ac line!) down to 0.5, 1, and 2 Hz for calibration purposes. ■

References

1. Holsinger, W.P. and Kempner, K. M., "Portable EKG Telephone Transmitter," *IEEE Trans. Biomed. Eng.*, 19, pp. 321-323, 1972.
2. Klingenstein, C.H. et al, "A Method of Computer-Assisted Pacemaker Surveillance From a Patient's Home via Telephone," *Computers and Biomedical Research*, 6, pp. 327-335, 1973.

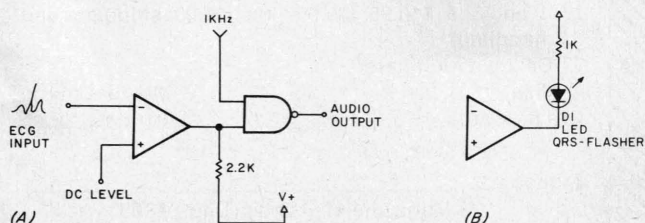
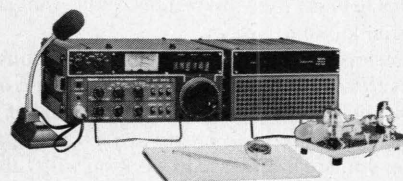


Fig. 6. (a) Qrs-beeper. (b) Qrs-flasher.

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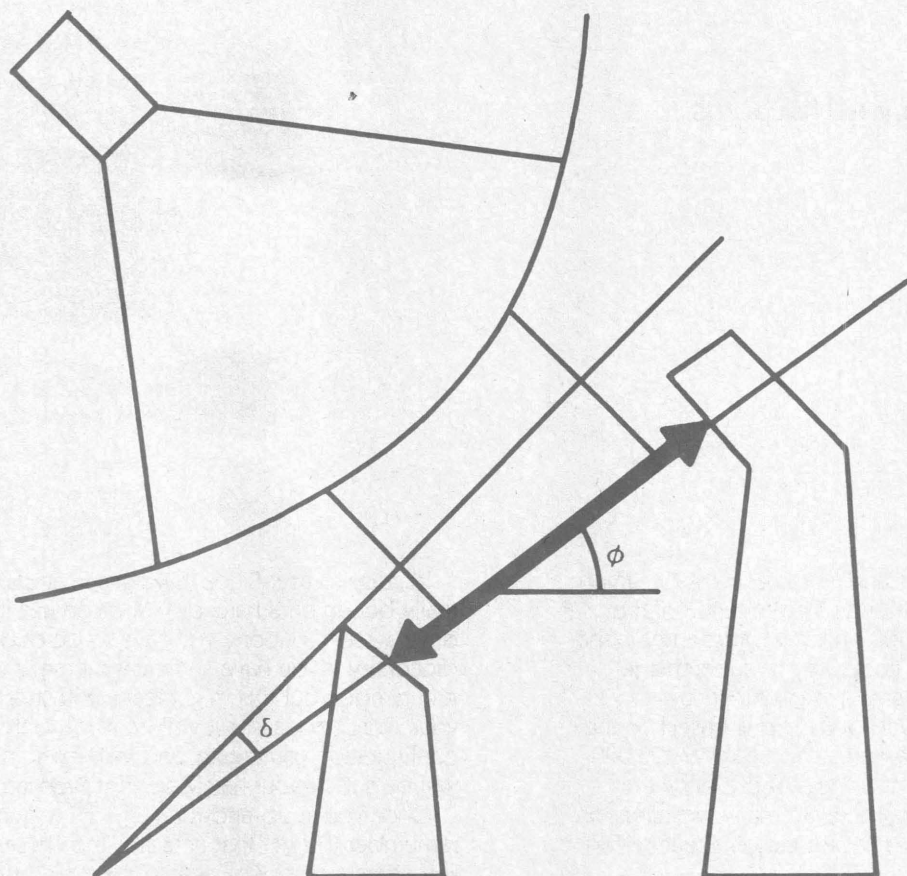
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TVRO ANTENNA POINTER PROGRAM



Your computer can accurately point your TVRO antenna.

Edmund T. Tyson

■ Most articles give pointing angles in azimuth/elevation coordinates. But the working coordinate system for most antennas is identical with the astronomical hour angle/declination coordinate system. In this system, the hour angle axis is parallel with the spin axis of the Earth and the declination axis is orthogonal to the hour angle axis. So the antenna points to any satellite in the Clarke belt by motion about a single axis. The Clarke belt is a band in space above the equator, some 22,240 miles up. Satellites in this belt make one revolution about the Earth in a day, so they appear fixed in one spot in the sky.

About the program

This program calculates angles for pointing TVRO antennas at synchronous satellites. It is written in BASIC for use on many computers. We tested it in midsummer by computing telescope settings to observe satellites broadcasting TV signals. A telescope observatory in the southwest confirmed the angles by seeking and finding the satellites. A dial was designed that permits pointing an antenna at a chosen satellite without searching.

The program is useful only for calculating synchronous satellite positions and is based on these assumptions: Inclination of the satellite orbit to the

```

100 ' COMPUTE POINTING ANGLES
105 '   FOR TVRO ANTENNA
110 '
120 DEFSNG A-H, 0-Z
130 DEFINT I-N
140 CLS
150 PI=3.141593#
160 D2R=PI/180!
170 RE=6378.165
180 '
200 ' **** OBSERVER COORD ****
210 PRINT "OBSERVER N LAT/W LONG";
215 INPUT ,FLAT,FLONG
220 FLONG=-FLONG
221 WLONG=-FLONG
225 LPRINT "TVRO AT N LAT=";
226 LPRINT ,FLAT," W LONG=";WLONG
227 LPRINT
230 X=1!
240 AX=0!
250 Z=0!
260 ' ROTATE ABOUT "Y" BY LAT
270 AX=FLAT
280 IAX=2
290 GOSUB 1000
300 ' ROTATE ABOUT "Z" BY -LONG
310 AX=FLONG
320 IAX=3
330 GOSUB 1000
340 XO=-X
350 YO=-Y
360 ZO=-Z
390 LPRINT " SAT LONG AZIM";
391 LPRINT " ELEV HR AX";
392 LPRINT " DEC AX RANGE KM"
399 ' ****
400 ' **** SAT COORD ****
410 CLS
415 LPRINT
420 INPUT " SAT LONG ", SLONG
421 SLAT=0!
425 SLONG=-SLONG
426 GLO=-SLONG
428 ' RADIUS VECTOR OF
429 ' SYNC SAT = 6.611 ERU
430 X=6.611
431 Y=0!
432 Z=0!
435 ' ROTATE RV ABOUT "Y" BY SLAT
436 AX=SLAT
437 IAX=2
438 GOSUB 1000
440 ' ROTATE RV ABOUT "Z" BY -SLONG
441 AX=SLONG
442 IAX=3
443 GOSUB 1000
450 ' TRANSLATE EQ COORD
451 ' TO OBSERVER
455 XS=X+XO
456 YS=Y+YO
457 ZS=Z+ZO
460 ' COMPUTE HOUR ANGLE
461 ' AND DECLINATION
470 GOSUB 2000
471 DAH=(AX/D2R)*FLONG
472 REQ=SQR(XS*XS+YS*YS)
473 DAD=ATN(ZS/REQ)/D2R
474 RS=SQR(REQ*REQ+ZS*ZS)*RE
500 ' **** ROTATE TO TOPOS ****
501 '
510 X=XS
511 Y=YS
512 Z=ZS
515 ' ROTATE TRANS COORD
516 ' ABOUT "Z" BY (LONG-90)
520 AX=-FLONG-90!
521 IAX=3
522 GOSUB 1000
530 ' ROTATE TRANS COORD
531 ' ABOUT "X" BY (LAT-90)
535 AX=FLAT-90!
536 IAX=1
537 GOSUB 1000
540 XS=-Y
541 YS=X
542 ZS=Z
543 RT=SQR(XS*XS+YS*YS)
545 GOSUB 2000
550 AZ=AX/D2R
551 EL=ATN(ZS/RT)/D2R
LPRINT USING"*****";GLO;
LPRINT USING"*****";AZ;
LPRINT USING"*****";EL;
LPRINT USING"*****";DAH;
LPRINT USING"*****";DAD;
LPRINT USING"*****";RS
580 GOTO 400
999 ' *****
1000 ' COORD ROTATION ROUTINES
1010 '
1020 AXR=AX*D2R
1030 SINX=SIN(AXR)
1040 COSX=COS(AXR)
1050 XT=X
1060 YT=Y
1070 ZT=Z
1080 ON IAX GOSUB 1100,1200,1300
1090 RETURN
1100 ' ****X ROT****
1110 X=XT
1120 Y=YT*COSX+ZT*SINX
1130 Z=-YT*SINX+ZT*COSX
1140 RETURN
1200 ' ****Y ROT****
1210 X=XT*COSX-ZT*SINX
1220 Y=YT
1230 Z=XT*SINX+ZT*COSX
1240 RETURN
1300 ' ****Z ROT****
1310 X=XT*COSX+YT*SINX
1320 Y=-XT*SINX+YT*COSX
1330 Z=ZT
1340 RETURN
2000 '
2005 ' QATAN
2010 ' ARC TAN IN PROPER QUADRANT
2020 IF XS<0 GOTO 2060
2030 IF YS=0 THEN AX=0: RETURN
2040 IF YS>0 THEN AX=PI/2: RETURN
2050 IF YS<0 THEN AX=1.5*PI: RETURN
2060 AX=ATN(YS/XS)
2070 IF XS<0 THEN AX=AX+PI: RETURN
2080 IF YS<0 THEN AX=2*PI+AX
2090 RETURN

```

equator is small, eccentricity of the orbit is small, and the semi-major axis of the orbit is 6.611 Earth radii.

You need north latitude and west longitude of the antenna and west longitude of the satellite. South latitude and east longitude values must be entered as negatives. The program can be used for any geographical location of the observer and satellite longitude. Negative elevation angles show the satellite below the horizon for that location.

When executed, the program prompts for the geodetic position of the antenna and this is printed to label that data. Antenna position is then converted to geocentric rectangular coordinates and saved. Additional prompts request west longitude of each new satellite. These positions are converted to geocentric rectangular coordinates, antenna coordinates are subtracted and the vector is solved for hour angle and declination.

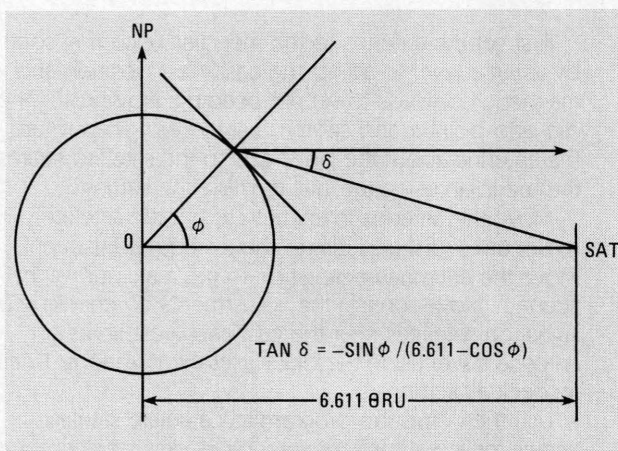


FIG. 1—DECLINATION OR OFFSET ANGLE is shown above. This is a sketch of Earth and satellite seen from the equatorial plane. The relation is described by the equation.

The hour angle

Hour angle is measured about the polar axis of the antenna mount. It's zero when the satellite is on the meridian, negative when the satellite is east of the observer, positive when it is west. Declination is the angle created by the observer's displacement from the equator. It is included in the mount setting by the *offset*

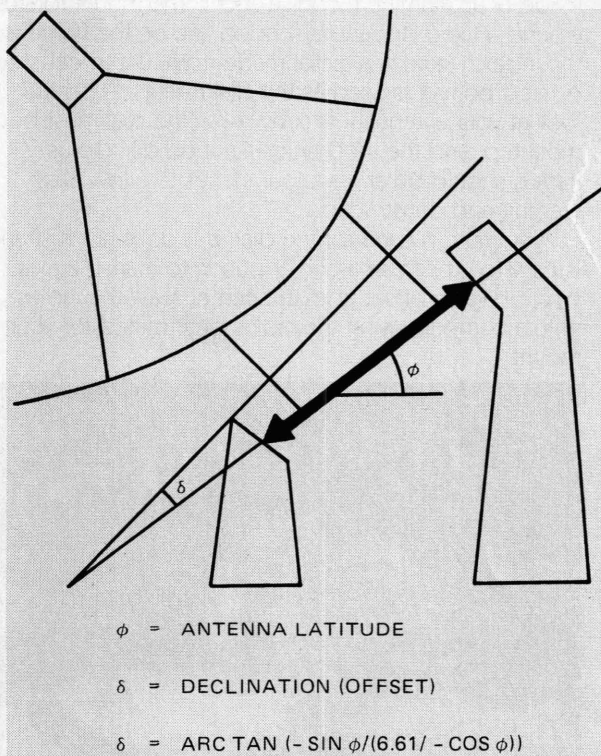


FIG. 2—HOW LATITUDE and declination angles are included in the antenna mount. The difference vector is rotated to the observer's position and solved for azimuth, elevation and slant range.

adjustment made during mount and dish assembly. Figure 1 shows the declination or offset angle. This is a sketch of Earth and satellite seen from the equatorial plane. Distance from satellite to Earth's center is 6.611 Earth radius units. An increase in latitude causes a related increase in declination. The equation describes the relation. Once calculated, a fixed angle for declination can be used as the change in declination from a meridian to an extreme east or west position is small compared to antenna beamwidth at C-band. Figure 2 shows how latitude and declination angles are included in the antenna mount.

The difference vector is then rotated to the observer's position and solved for azimuth, elevation and slant range. Azimuth is in degrees from north around through east as in the graduations of a compass. Elevation is in degrees above the horizon with zero degrees at the horizon and 90 degrees directly overhead. Slant range is the distance from antenna to satellite.

The printed output lists observer's position, satellite longitude, azimuth, elevation, hour angle, declination angle, and slant range in kilometers. Almost all TVRO antenna mounts use the hour angle/declination mode of motion. This is the best mount since it lets the antenna be aimed at any satellite by moving it on a single axis. This also allows constructing a simple dial that lets you point at any interesting satellite.

The dial

Figures 3 and 4 are typical dials. Figure 3 is a dial made of an aluminum cookie sheet. The mount rotates around a fixed dial and a scribed line on the frame is the index. Figure 4 is a dial made from .040 sheet stock. A fixed pointer is used as the dial rotates. A careful look at your antenna will show how the dial must be mounted, and the mechanical limit on dial radius. Radius should be at least four inches to allow easy reading and construction.

Aluminum is used for the dial as it does not rust and is easy to work. Use a center punch to mark the axis which can be offset from the center. Mark the radius with a compass using the radius determined for your mount.



FIG. 3—AN ALUMINUM COOKIE SHEET is used as a dial. The antenna mount rotates about a fixed dial with a scribed line on the frame as an index.

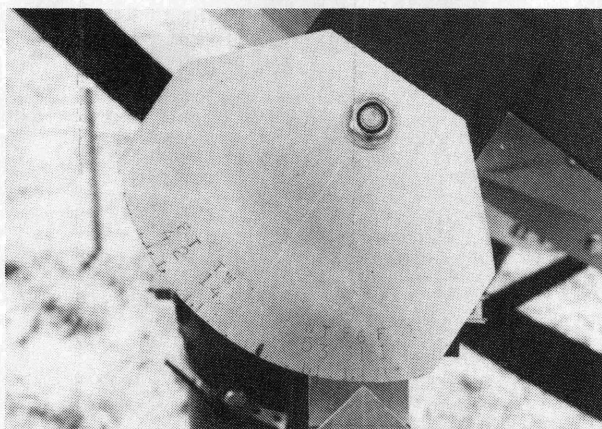


FIG. 4—DIAL FORMED FROM ALUMINUM STOCK. A fixed pointer is used since the dial rotates. Look at your own antenna to select the best form.

With a straightedge, divide the arc in halves to represent meridian or zero point and scribe a line through the punch mark and arc. A protractor is used to mark five degree angular increments of satellite positions as calculated by the computer program. Make sure you make the marks on the correct half of the dial. Satellites to the east of the antenna are prefixed with a negative (-) sign. The correct half changes depending on whether the dial rotates or is stationary.

Angular divisions of five degrees are adequate and can be marked by a double punch mark. Individual satellites are identified by scribing a mark at calculated angles along with an alpha-numeric label such as F4, W4, G1, etc. Antennas separated by a few miles need different dials as a one-degree shift in longitude results in at least a one-degree change on the dial.

A dial is a useful tool for technicians installing many antennas. The time to align can be reduced and precision improved. A dial calculated for the area where installations are done and marked with the east, west and meridian satellites will speed the process of alignment with the Clarke belt and setting remote positioning actuators and indicator systems.

Dial alignment procedure

First set the antenna to the meridian (pointing south) by using a level to adjust the east/west position and a magnetic compass to set the heading. Now install the dial and pointer and set the dial to zero. Then rotate the antenna about the polar axis to the satellite nearest the meridian using the dial to make the setting.

Move the antenna in azimuth to get the satellite with maximum signal and clamp the azimuth adjustment. Move the antenna in elevation to get maximum signal strength. Now confirm the adjustments by checking the signal on satellites near the east and west limits of antenna travel using the dial to set the positions. Finally, recheck all settings.

Use the computer program to calculate satellite angles for a dial for your antenna. A dial is an essential accessory for manually positioned antennas. It will eliminate guesswork when changing from one satellite to another. ◀▶

A home or small-business intercom needn't be expensive.

Our economical system uses standard pulse-dial telephones, and has options for a PA system—with music!



NINE-STATION INTERCOM

DWIGHT MORRISON

INSTALLING A BUSINESS-TELEPHONE system can be an expensive proposition. But if you have several unused pulse-dial telephones lying around, you can use them and cut costs dramatically by building our simple control center (for about \$50). The control center provides a switchboard-like function allowing any one of nine phones to call any other. In addition, one station can be set up as the office PA system; to make an announcement from any station, just dial the station assigned to the PA. The PA system can provide background music (from any source) while no announcement is being made.

System features

Nearly any telephone with a pulse-dial output may be connected to the control center. For example, a standard rotary-dial telephone, a pushbutton phone, a speakerphone, or even a cordless phone could be used. (See this story's lead photo.) Each telephone may be located as far as 2000 feet from the control center; interconnections are made with standard four-conductor 22-gauge telephone wire.

Standard telephone-ring generators require a special 90-volt, 20-Hz power supply. Building such a circuit is expensive as well as difficult because components are hard to obtain. Therefore, the control center generates its own special ring and dial tones.

The control center provides a LIGHT output to illuminate a "busy" LED installed in each station. All LED's light up whenever any station is off hook.

How to use it

Operating the intercom is simple. Pick up the phone at any station; you'll hear a

dial tone in the earpiece, and all LED's will light. Now you can dial any station; while dialing occurs, all of the LED's will flash. After dialing, you'll be able to hear the ringback tone in the earpiece. In addition, the station you dialed will buzz for about one second. If no one answers at that station, you can call again without hanging up.

If someone at a different station picks up his phone, he can join in the conversation. That "party-line" effect can be used for simulating a conference call. For example, if two people are talking and decide that they'd like to include a third, they should hang up, and one should call the third. After giving him time to answer, the one who hung up can pick up his phone and join the conversation.

Circuit operation

The schematic diagram of the control center is shown in Fig. 1. The heart of the circuit is an M-959 IC, manufactured by Teltone (P. O. Box 657, 10801-120th Avenue N. E., Kirkland, WA 98033-0657). The M-959 is a CMOS device that counts dial pulses and provides an encoded binary representation of those pulses. For example, if the digit nine were dialed, the D0-D3 outputs would contain logic levels 10 01. The M-959 also provides a logic-level indication of hook status (OH) at pin 13.

When any station goes off hook, current flows through the coils of relays RY10 and RY11, so their contacts close. Relay RY10 supplies power to the LED's in each station, and relay RY11 grounds the LC (Loop Current) input of IC1. That forces the OH output to go high.

The SR flip-flop composed of IC4-c and IC4-d does not change state, but the

astable oscillator composed of IC5-a and IC5-b turns on because both OH and pin 10 of IC4-c are high. That astable is what provides the dial tone, which is fed to the talk circuit via R12 and C12.

When a digit is dialed, relays RY10 and RY11 "follow" the dial pulses—i. e., their contacts make and break a number of times according to the digit that was dialed. After dialing stops, IC1 places the encoded binary digit on pins 8–11. In addition, IC1's STB output (pin 12) goes high for 200 ms.

There is no station one in this system. The reason is that the circuit cannot distinguish well between going off hook and dialing the digit one. The four gates of IC3 are used to ensure that only stations greater than one are called. Those gates are set up as a three-input OR gate. The inputs of the gate are connected to the D1-D3 outputs of IC1, so any station greater than one will cause the output of IC3-d to go high. That signal is NAND-ed with the STB output of IC1 by IC4-a, and the combined signal is used to trigger IC6, a 555 timer operated in the one-shot mode.

With the component values shown in Fig. 1, the output of the 555 will remain high for about one second. The output of IC4-a also resets the IC4-c/IC4-d flip-flop, thereby disabling the ring generator (IC5-a and IC5-b).

The 555's output (pin 3) enables the astable composed of IC5-c and IC5-d, which oscillates at a frequency of about 1000 Hz. That astable supplies the dial tone, which is fed to the talk circuit via R15 and C12.

In addition, the 555's output is inverted by IC4-b and that signal is used to strobe the binary outputs of IC1 into IC2, a 4-

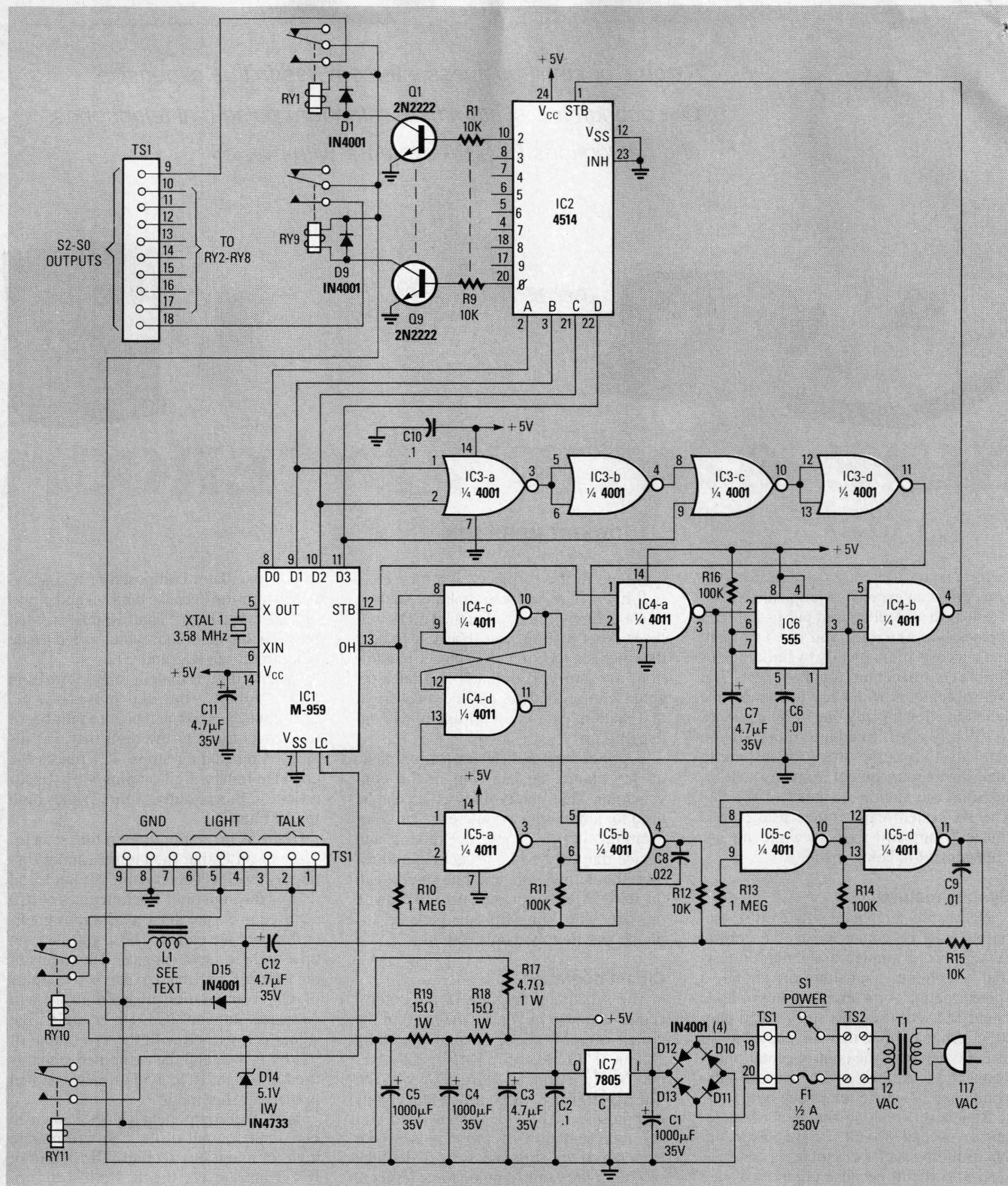


FIG. 1—A-HALF-DOZEN IC'S and a dozen relays comprise most of the circuitry of the intercom. Note that only two of the relay-output circuits are shown (R1-Q1-D1-RY1 and R9-Q9-D9-RY9); the other seven outputs from IC2 are wired in a similar manner.

to-16 line decoder that works as follows. Assume that station two was dialed. Then pin 10 of IC2 will go high and turn on transistor Q1, which will in turn enable relay RY1. At that point, 16 volts will be present at pin 9, the S2 output, of TS1, the 20-terminal strip. That voltage would then

drive the buzzer in station two.

When the conversation ends, both parties hang up. Then relays RY10 and RY11 will de-energize. That will cause all the LED's in the circuit to extinguish, and it will allow IC1's LC input to float high.

Diodes D10-D13 rectify the 12-volt AC

input; IC7, a 7805 regulator provides +5-volts for the logic IC's. Resistors R17, R18, and R19 provide +16-volts to operate the relays and the buzzers. Coil L1, which is actually the primary of a 500-ohm audio transformer, filters power-supply hum from the talk circuit.

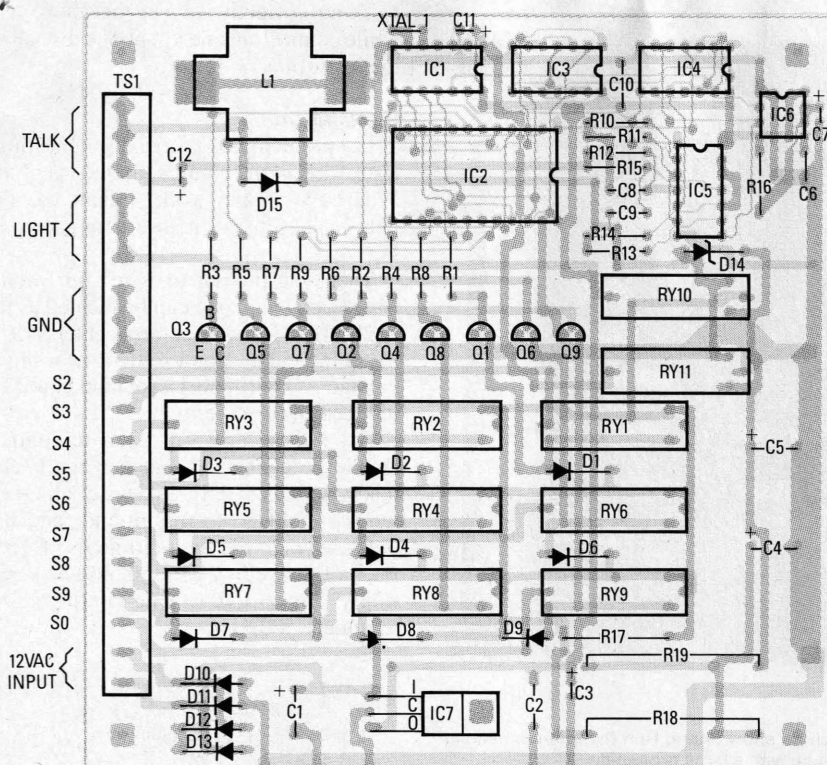


FIG. 2—MOUNT ALL COMPONENTS as shown here. The terminal strip (TS1) is composed of ten dual terminal strips.

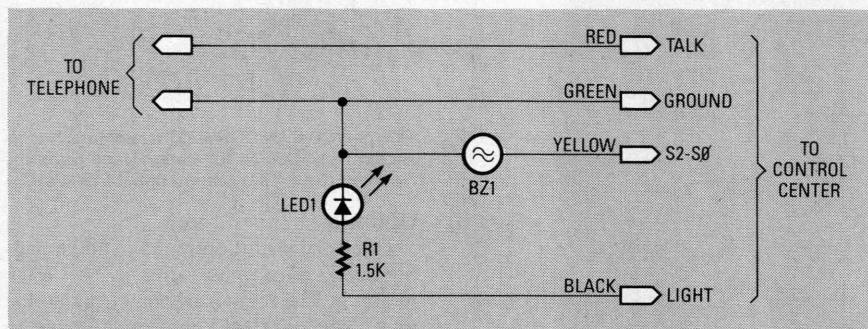


FIG. 3—AN LED, A BUZZER, and a resistor should be mounted in each telephone.

Assembly

Terminal strip TS1 is composed of ten terminal-strip pairs; each PC-board pin and screw terminal is located 0.2" from its neighbor. Radio Shack and Digi-key (P. O. Box 677, Thief River Falls, MN 56701) sell connectors from different manufacturers that will fit the board.

Foil patterns for etching a suitable PC board are shown in PC Service. Beware that the board is double-sided, so, if you etch your own board, you'll have to make provision for soldering all component leads (about 200) on both sides of the board. Alternatively, you can buy a board from the source mentioned in the Parts List. The commercially available board has plated-through holes.

Whether you buy or build your PC board, check it carefully for shorted and open traces before mounting any components. Then, after solving any problems, solder the components to the board ac-

cording to Fig. 2. Mount the low-profile components and IC sockets first.

Don't insert the IC's in their sockets until you perform the initial check-out. The exception is IC7 (the 7805); bend its legs 90° so that it rests flat against the PC board. Secure its tab to the board.

Lay the PC board on a non-conductive surface and connect a 12-volt AC source to the board through a ½-amp fuse. Measure the voltage at the input terminal of IC7; it should be about 16-volts DC. The output of IC7 should be between 4.8- and 5.2-volts DC. If those voltages are correct, remove power and insert the IC's in the appropriate sockets.

Testing

You'll need to add a buzzer, an LED, and a resistor to (at least) two telephone sets as shown in Fig. 3. Four wires connect each phone to the control center: three for the talk, light, and ground circuits, and one from each phone to the

PARTS LIST

All resistors are ¼-watt 5% unless otherwise noted.

R1-R9, R12, R15—10,000 ohms
R10, R13—1 megohm
R11, R14, R16—100,000 ohms
R17—4.7 ohms, 1 watt
R18, R19—15 ohms, 1 watt

Capacitors

C1, C4, C5—1000 µF, 35 volts, electrolytic
C2, C10—0.1 µF, ceramic disk
C3, C7, C11, C12—4.7 µF, 35 volts, electrolytic
C6, C9—0.01 µF, ceramic disk
C8—0.022 µF, ceramic disk

Semiconductors

IC1—M-959 dial-pulse counter (Teltone)
IC2—4514 4-to-16 line decoder
IC3—4001 quad NOR gate
IC4, IC5—4011 quad NAND gate
IC6—555 timer
IC7—7805 5-volt regulator
D1-D13, D15—1N4001 rectifier
D14—1N4733 5.1-volt, 1-watt Zener diode
Q1-Q9—2N2222

Other components

F1—0.5 amp, 250 volts
RY1-RY9—reed relay, 12 volt
RY10, RY11—reed relay, 5 volt
L1—500-ohm audio transformer (see text)
S1—SPST toggle
T1—12-volt, 0.5-amp wall transformer
TS1—20-position terminal strip (see text)
TS2—2-position barrier block
XTAL—3.58 MHz, color-burst

Miscellaneous: Enclosure, fuse holder, telephone wire, 12-volt piezo-electric buzzers, LED's, and resistors for telephone stations, etc.

Ordering info

Note: The following are available from COM-TECH, 1856 S. Highland, Jackson, TN 38301: PC board, IC1 (M-959), IC2, and L1, \$34.95; 12-volt, 0.5-amp wall transformer, \$11.95; 12-volt DC latching relay with 3-amp contacts, \$19.95. All orders add \$3 (\$5 Canada) for shipping. Foreign orders must include U. S. funds and \$8.00 for shipping. Tennessee residents add 7% sales tax.

proper S2-SØ output. At the phone end, the red and green wires going to the telephone should connect in parallel with the corresponding wires already in the telephone. The other two wires are connected to the LED and the buzzer as shown.

With power disconnected, connect two telephones to the control center. Apply power, and lift the handset from the base of one telephone. All the busy LED's should light up, and a dial tone should be heard through the handset. Dial the number corresponding to the other telephone. The busy LED's should flash during dialing. After dialing, a ringback tone should be heard over the handset, and the other phone should buzz, both for about one second. Lift the handset from the other phone, if all is well, you should be able to communicate over the two handsets.

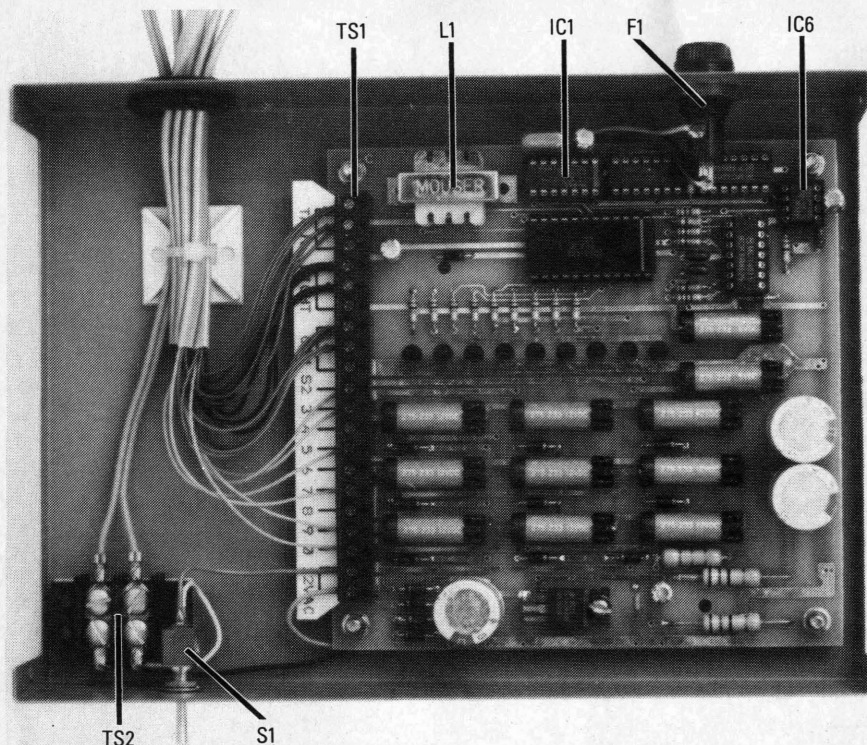


FIG. 4—MOUNT THE PC BOARD, fuse, and power switch as shown here. Run the interconnecting wires through a hole in the rear of the box; insulate the hole with a large grommet.

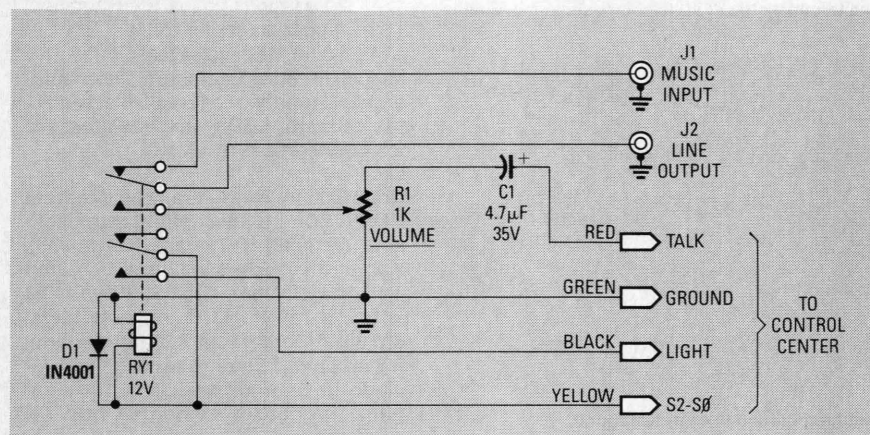


FIG. 5—ADD A PAGING/MUSIC circuit as shown here. Connect the music input to the output of a transistor radio or another suitable source.

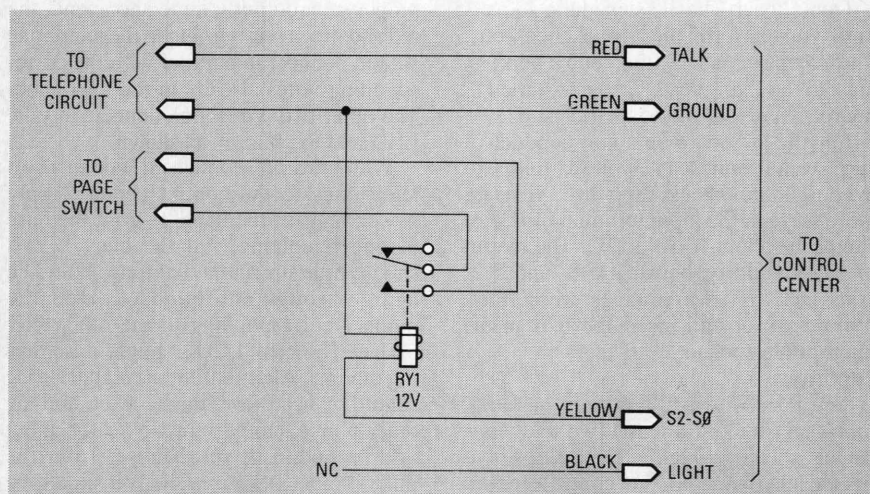


FIG. 6—CONNECT A WIRELESS PHONE as shown here.

To verify that all other channels are working correctly, repeat the above test while connecting one test phone to each of the S2-SØ pins of TS1.

Installation

If a problem is detected during testing, track down the source and correct it. When all circuits work, mount the PC board in a suitable enclosure, as shown in Fig. 4.

Then you'll want to wire your premises. Each telephone can be located as far as 2000 feet from the control center. When installing cable, don't route it near high-voltage AC wiring because hum could be induced in the system.

All red wires from each telephone should be connected together and to the talk terminals of TS1. Likewise, all green wires should go to the ground, and the black wires to the light terminals of TS1. Each of the yellow wires should be separately connected to one of the S2-SØ terminals of TS1.

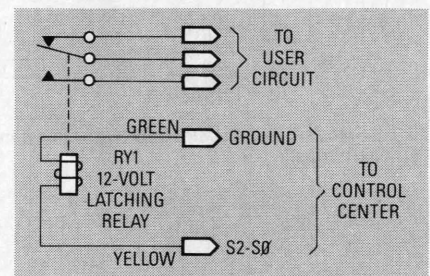


FIG. 7—YOU CAN CONNECT a latching relay as shown here. Dialing its station once turns the relay on; dialing it a second time turns it off.

Options

One very useful option is to add a paging/background music circuit, as shown in Fig. 5. The relay is normally off, so the music source (J1) is connected directly to the line output (J2) through the upper set of contacts. But when that station is called, the relay activates, and the talk circuit is connected to the line output.

The S2-SØ outputs of the control center can be connected to devices other than telephones. For example, you could use one output to activate the "page" feature of a cordless phone. You could page someone carrying a cordless phone by connecting the paging switch in the base station to the control center as shown in Fig. 6.

Another option is to connect a latching relay to one control center output as shown in Fig. 7. Each time the station that circuit is connected to is called, the output of the relay would change state.

There are many other uses for the circuit. For example, you could control the latching relay from a cordless telephone to provide remote control of an electric garage-door opener. Many other uses will doubtless occur to you as you ponder the possibilities!

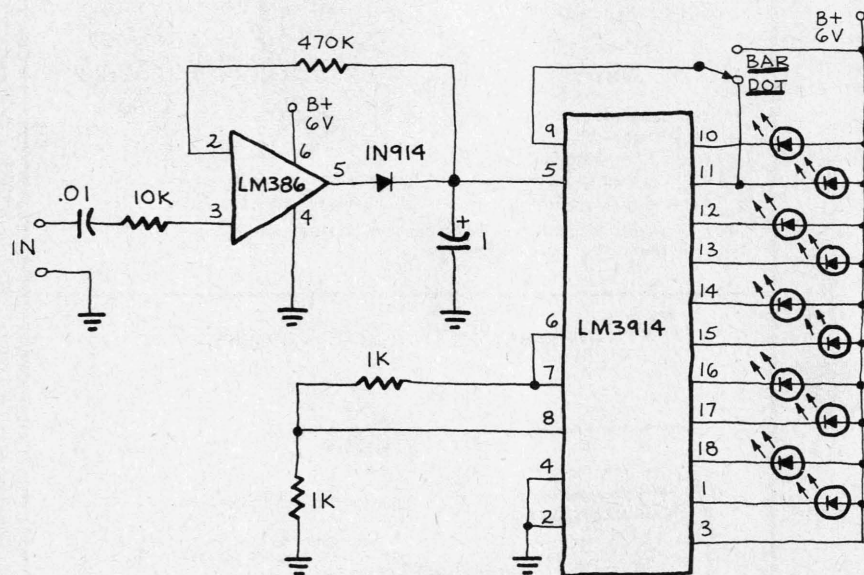
R-E



LED PEAKMETER

I WOULD LIKE TO SUBMIT MY LATEST project to your New Ideas column. I call it the LED Peakmeter. It is a basic dot/bar readout built around the new National LM3914 display driver. The circuit also includes a peak detector that immediately drives the readout to any new higher signal level and slowly lowers it after the signal drops to zero. The readout is a moving dot or expanding bar display.

The diagram shows one channel of the stereo LED Peakmeter shown in the photograph. All parts are easily obtained and layout is not at all critical. Although not absolutely necessary, I suggest trying the



circuit on a solderless breadboard before hand-wiring to check delay time and to match components in a stereo unit.

I used a spare piece of perforated board as a template to drill holes for the LED's in the project box's plastic front. A battery holder with four "C" cells is mounted on the back of the box.

The circuit has other possibilities. It can be expanded for a longer bar readout if desired. Tapping five or more LED Peakmeters into a frequency equalizer or series of audio filters should give a unique result. Physical layout of the LED's can also be changed to simulate the action of regular VU meters.

The bottom LED of each peakmeter remains on with no signal at the input, thus providing a pilot light for the unit.—
Wm. J. Cikas

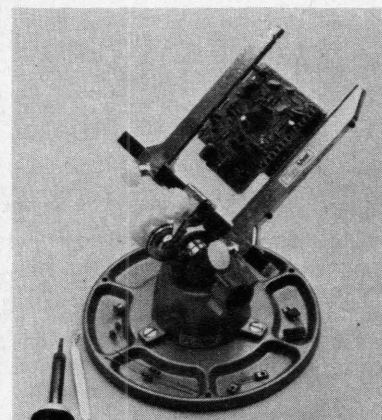


"So, we'd save 25 cents on the steak here. Now, how about the drinks?"

NEW IDEAS

This column is devoted to new ideas, circuits, device applications, construction techniques, helpful hints, etc.

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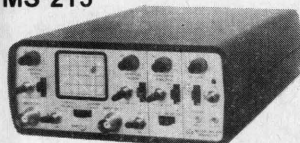
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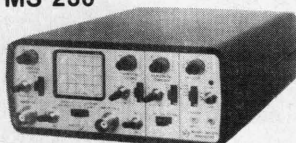
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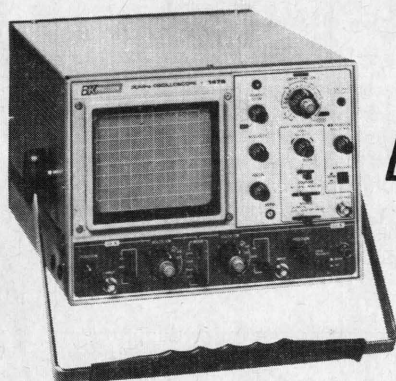
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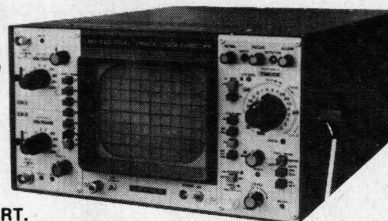
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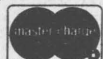
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BACKYARD SATELLITE TV RECEIVER

Part 6: The front end is critical if you build your own satellite TV receiver. This issue we explore several different approaches to making one that will work.

ROBERT B. COOPER, JR.

LAST MONTH, THE BASIC DO-IT-YOURSELF satellite TV receiving system was described along with a novel spherical antenna system. This month, we'll look at several approaches to building the front end of the receiver.

Suitable LNA designs

The low-noise amplifier (LNA) decision depends largely on the mixing approach taken by the builder. As discussed last month, if you decide to use a prepackaged passive double-balanced mixer, such as the VARI-L DBM 500 unit, you will need more voltage gain from the LNA than if you elect to use an active GaAs-FET mixer. We'll show both LNA approaches here: the bipolar transistor system for use where 40 to 50 dB of gain is required, and the GaAs-FET transistor system where approximately half as much gain is needed.

A few comments are in order for those building microwave circuits for the first time. Read them carefully.

1. **Board material**—Normal circuit-board materials, such as the familiar G-10, are bad news at microwave frequencies. Any printed-circuit board *must* be designed for microwave applications. That means a microwave-rated Teflon dielectric board. Such board material is expensive but if you use very small amounts of it, the per-system costs will still be minimal.
2. **Double-sided**—Use only double-sided board for all circuits, including those at baseband fre-

quencies. IC and packaged active devices used in this system, even when operating at baseband (video) frequencies, will oscillate when given the opportunity. (One recommended source for the microwave region board material that is used in the 4 GHz LNA stages and in the local oscillator/active mixer segments is the Rogers Corporation, Box 700, Chandler, AZ 85224. The board material is Duroid grade D-5880 226-127; dielectric thickness is 0.031 inches, 1 ounce clad on two sides.)

3. **Grounds**—All boards must be perimeter-grounded. That means all around, all four edges, both sides. Spot grounds through standup mounting lugs or pillars are not adequate.
4. **Lead length**—Exceedingly short, direct leads must be used with all parts. Remember that at microwave frequencies even a 1/8th-inch lead becomes an appreciable portion of a wavelength.
5. **Capacitors**—All capacitors specified in the microwave portion must be chip type. Normal ceramic, etc. capacitors have far too much inductance at microwave frequencies to be utilized. Where RF chokes are specified, put them in.

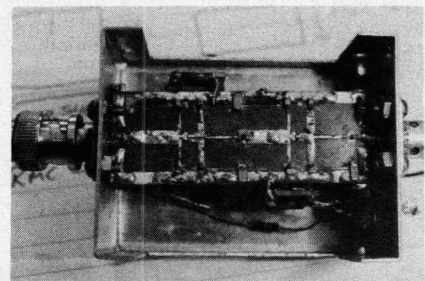
There are several sources for chip capacitors suitable for this project. One national source is Dielectric Labs, 69 Albany St., Cazenovia, N.Y. 13035. Smaller quantities can be obtained from Robert M. Coleman, RFD 3, Box

58-A Travelers Rest, SC 29690, and, from Satellite Innovations, Box 5673, Winston Salem, NC 27103. Where some of the circuits here specify certain brands of parts, such as capacitors, look to the *value* of the device and then locate a suitable substitute from the sources just given.

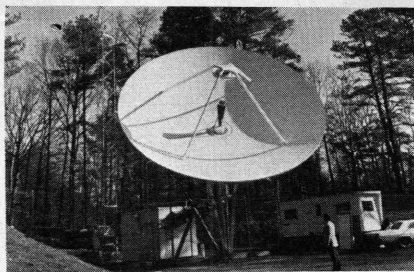
Two-stage bipolar LNA

The workhorse amplifier in this service is described in *Hewlett-Packard Applications Note 967*: a single-stage bipolar amplifier using either the HXTR-6102 or the HXTR-6101 devices. The 6102 is a better grade of the 6101 and it is capable of producing an LNA stage with approximately 10-11 dB of voltage gain in the 3.7 to 4.2 GHz range with a noise-temperature of between 270° and 290° Kelvin (K). The 6101 tends to be 15° to 25° K "hotter." (In this case, hotter is worse, not better!)

English experimenter Steve Birkill



TWO-STAGE BIPOLAR LNA is primarily an etched circuit board with very tiny parts mounted in precise position.



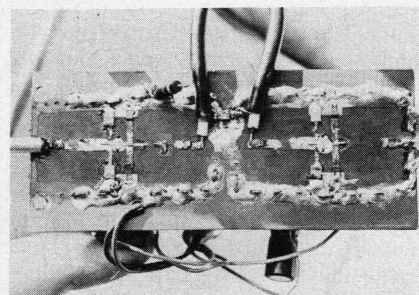
PARTS LIST

Two-stage bipolar LNA

- Q1—HXTR-6102 transistor (Hewlett-Packard)
 Q2—HXTR-6101 transistor (Hewlett-Packard)
 R1, R3—10,000 ohms, linear pot
 R2, R4—10,000 ohms, 1/2 watt
 C1, C8, C15—2.2 pF (Vitraron VJ0805A2R2DF)
 C2, C5, C9, C12—270 pF (Vitraron VJ0805A271)
 C3, C6, C10, C13—4.7 pF (Vitraron VJ0805A4R7DF)
 C4, C7, C11, C14—1000 pF (Vitraron VJ0805X102KF)
 PL1—SMA-type plug receptacle, tab contact, flush dielectric. Selectro type 50-646-4575-31 (gold plated) or similar
 J1—SMA-type jack receptacle, tab contact, flush dielectric. Selectro type 50-645-4575-31 (gold plated) or similar.
 Note: SMA connectors from different makers may be known variously as SMA, SRM, RIM or OSM.

Microstrip board: 62.5 × 22.5 × 0.79 mm.
 Duroid D-5880 226-127
 dielectric constant 2.5, etched.

of Sheffield has developed a two-stage circuit board using this device series and it is shown in Fig. 1. A full-size circuit board is shown in Fig. 2. The opposite side of the board—which, as a



TWO-STAGE HFET GaAs-FET LNA is similar in design and almost identical in layout to bipolar LNA two-stage device; primary difference being the substitution of HFET series transistors for bipolars.

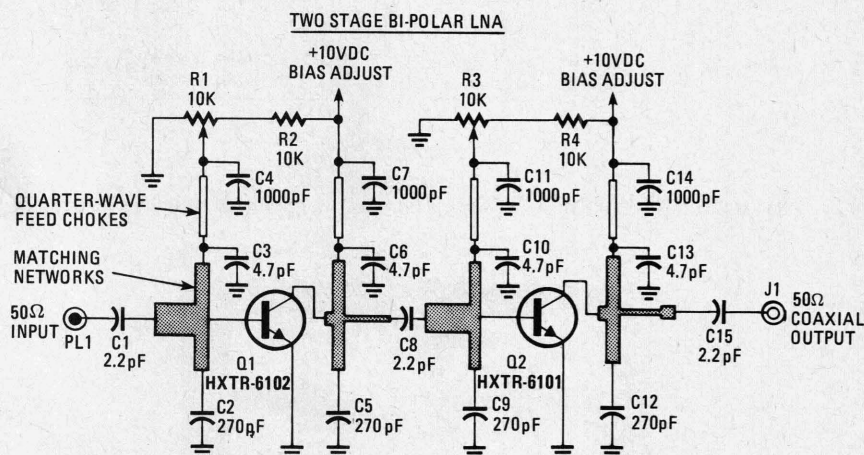


FIG. 1—SCHEMATIC DIAGRAM of the two-stage bipolar low-noise amplifier. The shaded areas represent leads and inductors that are vital parts of the circuit design.

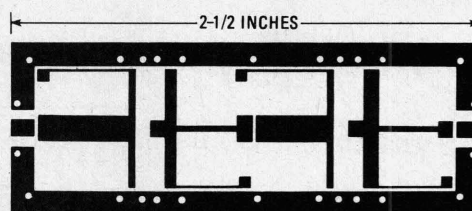


FIG. 2—PATTERN for etching the top surface of the LNA microstrip board. The lower surface is plain copper.

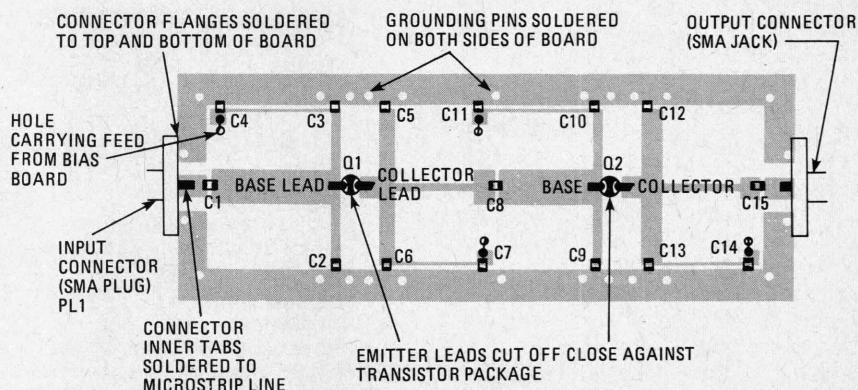


FIG. 3—COMPONENT LAYOUT shows placement of the transistors and capacitors. The capacitors are chip-type approximately 8 × 5 mm.

reminder, *must* be a microwave-rated board—is solid copper.

Following the components selection guide given here and the construction tips, there is nothing to the system in the way of tuning or alignment. Ten VDC is the operating voltage; the base bias is adjusted with the 10K pots (one per stage) for a total device current of 4 mA. There is no tuning other than this; all resonant circuits are obtained

with the etched inductances and the fixed capacitances shown.

Figure 3 shows a parts layout for the same two-stage amplifier. The bias parts (resistor plus pot per stage) can be located on the backside of the amplifier circuit board. When constructed, the board(s) must be mounted in a suitable microwave enclosure with suitable grounds all around as noted. The amplifier is very stable, but not when

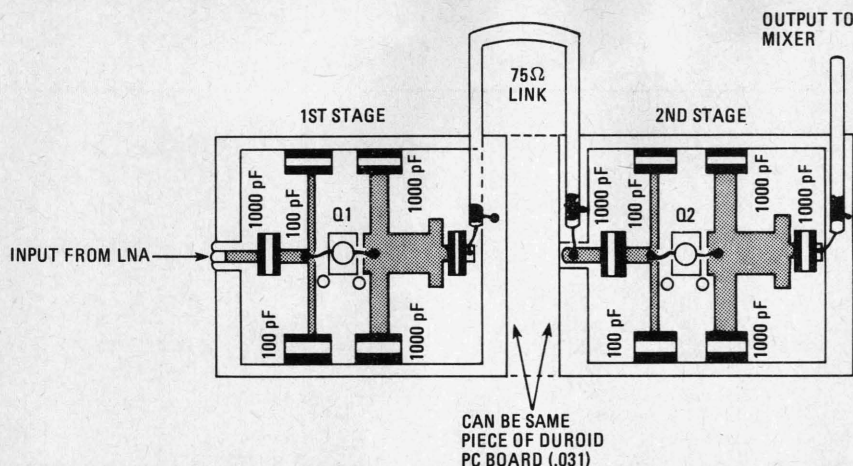


FIG. 4—SCHEMATIC AND LAYOUT of a two-stage LNA amplifier designed by Robert Coleman.

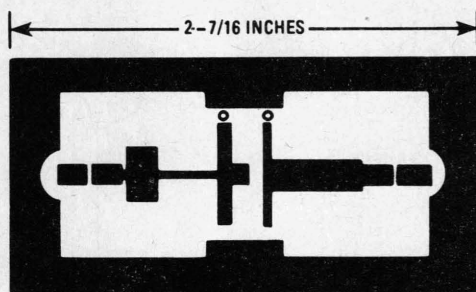


FIG. 5—PRINTED-CIRCUIT foil pattern for a single-stage low-noise amplifier. Two can be connected in cascade for more gain.

operating at the end of several clip leads as it dangles in space! One source for microwave enclosures is Adams Russell, Modpak Division, 80 Cambridge St., Burlington, MA 01803

Two-stage GaAs-FET LNA

If your approach is to follow the active mixer design of Robert Coleman, or you simply want a lower front-end noise figure than is possible with the HXTR bipolar series, then you can build the two-stage Coleman HFET-1101 amplifier. Figure 4 shows the parts layout for the HFET-1101 amplifier. The HFET series of GaAs-FET devices are also produced by Hewlett-Packard and a stocking distributor is Hallmark Electronics Corp., Attention: Paul Koeppen, 1208 Front St., Building K, Raleigh, NC 27609.

The HFET series of GaAs-FET's is capable of producing noise temperatures in the 170° K region (2-dB noise figure). Like the bipolar HXTR series, there is no tuning; the devices mount, turn on, and have voltage (positive and bias) supplies adjusted for optimum performance. Again, you cannot do that at the end of clip leads! The HFET data sheets suggest an operating voltage of +4.5 VDC. Developer Robert Coleman found that in the circuit shown (the actual-size foil pattern for a single stage is shown in Fig. 5.) the devices tended to be unstable at that voltage. By dropping the operating voltage to +3.6 VDC and applying a -3.0 VDC (adjustable) bias to the gate lead (as shown in Fig. 6) he was able to make the stage

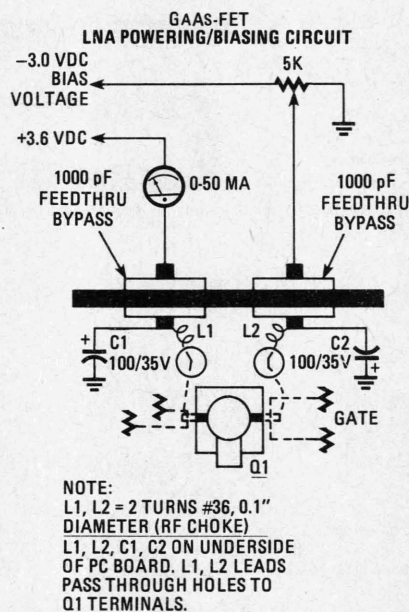
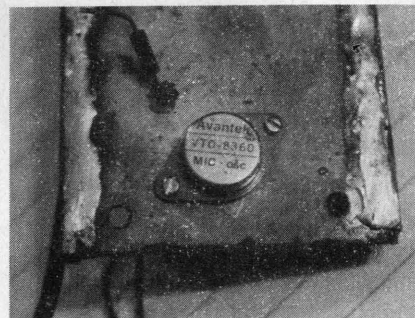


FIG. 6—HOW COLEMAN LNA IS BIASED AND POWERED. RF chokes L1 and L2 are mounted on underside of the board with leads anchored in holes in the PC board.

stable and optimize performance.

With all LNA stages (bipolar or GaAs-FET) there should be a separate bias control adjustment on each device. With the HFET devices, maximum gain occurs when the device current is around 40 mA but optimum noise figure occurs much lower; near 12 mA. Since in this situation voltage gain is secondary to noise-temperature performance, you will need a method of measuring the device current. Coleman's approach is to watch a current



THIS INNOCENT-LOOKING DEVICE is capable of producing +10 dBm of local oscillator signal at 4 GHz! Avantek VTO 8360 is a microwave oscillator device totally self contained. It mounts on full-foil side of board with pins (leads) accessible on opposite board side with active 4-GHz circuits.

meter on the stage and keep an eye on the satellite-delivered picture to optimize the stages involved. Start with the first stage after setting both stages to approximately 12 mA current.

Circuit boards are available for either the Birkill bipolar (two-stage) amplifier or the single stage GaAs-FET device from Robert M. Coleman, RFD 3, Box 58-A, Travelers Rest, SC 29690. The price is \$25 on the Birkill two-stage board and \$15 on the single-stage GaAs-FET board. A parts list is not included for the GaAs-FET LNA since many of parts are already listed for the bipolar LNA. The 100pF capacitors are also made by Vitramon and Q1 and Q2 are Hewlett-Packard HFET-1101 transistors.

The VTO local oscillator

Creating a +10 dBm-level continuous-wave signal source for the local oscillator can be a bit of a pain, especially when the local oscillator must operate in the 4-GHz region! Fortunately, Avantek (3175 Bowers Avenue, Santa Clara, Ca. 95051) has solved that problem with a neatly packaged device that only requires board mounting (on microwave pc board). The device requires connection of a +15 VDC supply and application of a second +10-to-+20-VDC range tuning voltage. The VTO 8360 device is virtually a perfect local oscillator source for our applications since it tunes the range of interest and while not inexpensive (in the \$125 region) it is far less costly to use than a lower-frequency oscillator chain with multiplying techniques. And, as Murphy notes, there is much less to go wrong because everything is inside on a substrate-designed package.

In Fig. 7 we have the complete local oscillator ready to drive any mixer put into service. The output pin four is linked through an appropriate short length of coaxial cable (if the length is under 6 inches, virtually any 50- or 75-ohm coax will function; but you will want to choose cable that will mate with the SMA or other series fittings you are using). Another approach is to

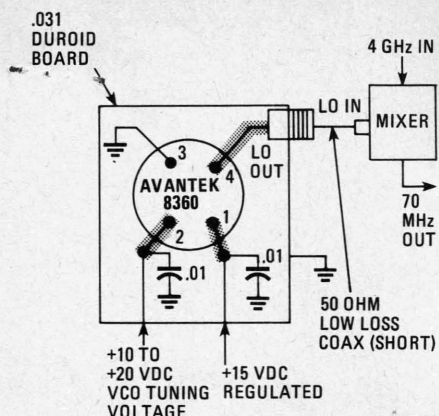


FIG. 7—THE AVANTEK VOLTAGE-CONTROLLED OSCILLATOR as it would be used as a local oscillator feeding a low-noise mixer.

use coaxial adapters to plug the output of the local oscillator directly into the appropriate input fitting on the mixer. If you are using the VARI-L DBM-500 mixer (VARI-L Company, Inc., 3883 Monaco Pkwy, Denver, CO 80297) you will need to build around the

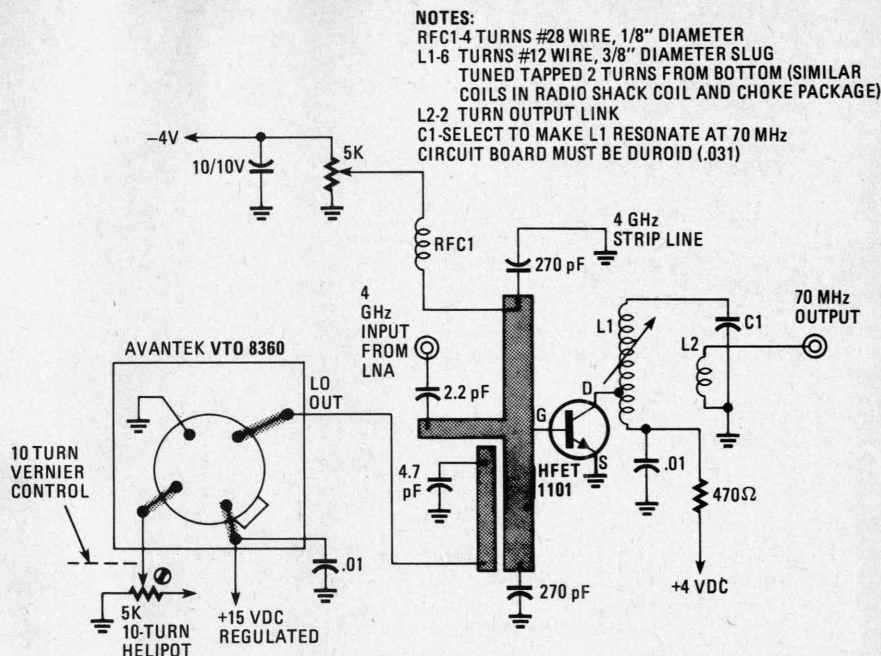


FIG. 8—A COMBINATION of the AvanteK VCO and GaAs-FET used to make a tuneable 4 GHz-10 70-MHz converter using an active mixer.

Additional Satellite Material

Satellite television reception enthusiasts interested in learning more about the fast developing satellite TV industry and the options available to persons building their own home terminals may find some of the following of interest:

1. **Satellite Study Package**—Designed to teach you how the satellite TV system operates, what the equipment requirements are, which services are available, and to whom and where. Includes a 72-page book written by Bob Cooper that explains in lay terms the complete satellite TV scene, plus a 22 x 35 inch four-color, two-sided wall chart depicting the location and operating characteristics of more than 30 geostationary satellites carrying television programming. Shipped via first class mail, price is \$15 in U.S. and Canada (in U.S. funds), \$20 elsewhere from: Satellite Television Technology, P.O. Box 2476, Napa, CA 94558.
2. **Coop's Satellite Digest**—A monthly publication providing up-to-date circuits, hardware, and satellite operational news. Mailed first class, widely read as the insider digest of the low-cost, private satellite TV industry. Price in U.S. and Canada is \$50 per year (\$75 outside, in U.S. funds); sample copy for \$5 in U.S. funds. Order from: Coop's Satellite Digest, P.O. Box G, Arcadia, OK 73007.
3. **Paul Shuch Satellite Lecture Series Videotapes**—Approximately eight hours in Beta or

VHS format; world-renowned microwave teacher and satellite system engineer-designer H. Paul Shuch takes the student through the entire satellite equation from antenna to remodulated RF. Series originally videotaped at SPTS '79, world's first international seminar for low-cost satellite TV terminals. Excellent learning tool, teaching tool. Price is \$210 in VHS (LP) and \$225 in BETA-2 in U.S. and Canada; add \$25 elsewhere from: Satellite Television Technology, P.O. Box G, Arcadia, OK 73007 (405-396-2574).

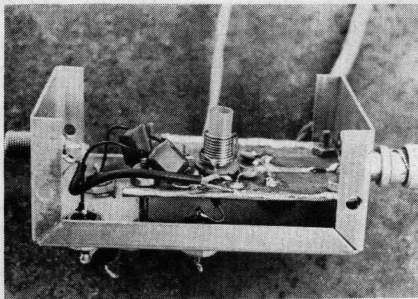
4. **SPTS '80/California**—A three-day lecture series and exhibit featuring noted satellite TV low-cost terminal-developers H. Taylor Howard of Stanford, Oliver Swan, who developed the Swan Spherical TVRO antenna, H. Paul Shuch of Microcomm, Robert Coleman of South Carolina, and many others. Combines classroom learning of the latest state of the art of satellite TV hardware, plus the latest in marketing of low-cost systems to private homes, with commercial exhibits of hardware. More than 25 sessions in three-day period with course learning materials. Next event will be held in San Francisco Bay Area in June of this year. For information, contact SPTS '80/California, P.O. Box G, Arcadia, OK 73007 (405-396-2574). Admission by pre-registration only, limited capacity.

“standard” microwave SMA fittings. Note that just as you don't use any substantial lengths of low-frequency (i.e., RG-8, etc.) coaxial cable at 4 GHz, you also don't use fittings such as the UHF type. Even the BNC type are at best questionable in performance at 4 GHz, although there are some type N fittings “rated” to beyond 4 GHz. The proper fittings and coax (for short interconnecting runs) can be located at Satellite Innovations, P.O. Box 5673, Winston Salem, NC 27103. What you are looking for is type SMA series connectors and suitable coax to mate with the SMA series fittings.

There is absolutely nothing to do with the VTO 8360 local oscillator but mount it and turn it on. The +10-to-+20-VDC tuning voltage varies the operating frequency through the range of interest (3.630 to 4.130 GHz). Once again—make sure the VTO 8360 is mounted on microwave circuit board and is firmly seated into a housing before turning on.

Active Mixer

The most cost-effective approach to the 4-GHz front-end at the moment appears to be a marriage of two stages of GaAs-FET LNA to the active mixer (plus local oscillator) shown in Fig. 8. This is another Robert Coleman-developed circuit, using the HFET-1101 not as an amplifier but rather as a single-ended mixer. The 4-GHz energy from the LNA stage(s) is coupled into the gate of the HFET 1101. The 4-GHz range local oscillator signal from the VTO 8360 is coupled into the same gate through a coupling strip. The 4-GHz pair of signals mix in the GaAs-FET



ACTIVE MIXER USING HFET device along with VTO 8360 (mounted out of sight on back side) mounts in single container. Unit can mount outside, at antenna, if appropriately weather-proofed thereby running only 70-MHz IF energy down and inside (in low-cost 50- or 75-ohm coaxial cable).

and are delivered at the output in the 70-MHz region. Inductor L1 plus capacitor C1 determine the IF resonance. With the value shown for L1, C1 will typically be around 5 pF. It is important that the Q of this output section be kept fairly low so that the full 30-MHz bandwidth of the 70-MHz IF signal gets out of the mixer and into the IF amplifier stages without being restricted. The 5K pot in the -4 VDC bias supply lead is adjusted for optimized performance simply by looking at the picture on the screen. This adjustment, plus the tuning voltage on the VTO 8360 are the only two real adjustments that you need to work with to get 4-GHz energy down to 70 MHz! Inductor L1 tunes broadly and can be optimized after the satellite signal is received.

This portion of the system can be tuned by using an MATV/CATV-type field-strength meter tuned in the 70-to-80 MHz region—or, in a pinch, you can actually run the 70-MHz IF output into a standard television receiver tuned to channel 4. No, you will not recover video (or audio); remember that the satellite TV format is FM, and 30-MHz or so wide FM at that, and consequently the 4.5-MHz wide TV IF set up to detect AM video modulation simply can't recover usable video. But, the TV receiver tuned to channel 4 can act as a "tuning indicator" of sorts, and if you happen to run across a transponder transmitting a static picture, such as color bars or a slide, you may for a brief instant even see something resembling a picture.

The circuit-board layout for the active mixer is available from Robert Coleman directly (address previously given) and a complete board ready to mount the parts on (including the VTO 8360) is also available for \$25.

There: Getting from 4-GHz down to the 70-MHz IF was not all that difficult! Next month we will look at the IF-to-baseband circuits, as well as the RF remodulation back to a standard NTSC format for direct viewing on a standard television receiver.

R-E

TELEVISION

WHAT'S 1980 TV

KARL SAVON
SEMICONDUCTOR EDITOR

HAVE YOU NOTICED THAT THE TELEVISION-receiver power transformer has virtually disappeared? Just two or three years ago one of the features of the more "solid" sets was the presence of that bulky, power-line isolating device. Today, design economy and a greater use of power-supply technology have eliminated the power transformer. Television tuners have also emerged, dramatically changed, from their mechanical infancy. Even many of the small-screen receivers use electronic tuners. The smaller sets tend to use the potentiometer-programmed varactor types first popular in the large-screen sets, while the larger deluxe sets now have "intelligent" tuning systems that smack of space-age technology and bear the fruits of microcomputer technology.

Those advances are found in both the surviving American producers' sets as well as the product releases of the Far Eastern competition.

Deflection and power-supply circuitry

Figure 1 shows the merged horizontal-deflection and power-supply circuitry of the 1980 Sharp 19D82 chassis. That receiver typifies the general circuit-design direction. The main chassis consists of four integrated circuits (two more are used in the tuning system) surrounded by a relative sparsity of discrete components.

The set's schematic displays an unusual neatness and simplicity for a color television receiver. Although the innards of the IC's themselves are shown in Fig. 1 as blocks, the schematic seems to lose many of the mysteries that were inherent in the esoteric, discrete designs of the past. It is no longer necessary for manufacturers to use every circuit trick possible to keep costs under control.

This particular deflection system uses a single IC that contains the sync separator, horizontal oscillator, vertical oscillator, high voltage hold-down (X-ray protection), and vertical preamplifier stages. There is no fundamentally new functionality in the deflection structure, but rather a new kind of organization that supports a SCR-based regulator system. The design eliminates the power transformer by transferring its responsibility to the horizontal-output transformer. In addition to the traditional pix-tube second-anode and focus supplies, the horizontal-output transformer drives the set's main 18-VDC low-voltage power supply through D704, as well as the regulated 110-VDC power supply.

As a result of the SCR regulator circuit, all supplies energized from the deflection transformer are regulated. One interesting thing is that the 110-VDC power supply feeds the horizontal-output transistor and so is self-supplied. It's not perpetual motion though, since all the energy ultimately comes from the 170-volt DC supply that runs from the AC

MEASURING RF POWER

Radio frequency (RF) power is one of the principal performance indicators of radio transmitters. Whether you are a ham-radio operator, citizen's-band operator, or a commercial radio-communications technician, the RF-power output of your transmitters is a key indicator of its health. RF power measurements vary from a few milliwatts up to many kilowatts. In this article we will take a look at some of the principal methods for measuring RF power.

RF power measurement is no different from other alternating-current (AC) power measurements, although the increased frequencies involved cause some problems. Power is still described by the standard equations ($P = V \times I$, $P = I^2 \times R$ and $P = V^2/R$), so if you can measure the current or voltage and know the resistance, you can measure power. The problem is relatively easy with sine-wave RF, such as unkeyed continuous wave (CW) signals, but becomes a bit more complicated on modulated, chopped, keyed, or pulse RF waveforms. In any event, the most basic power level measured is the root-mean-square (rms) value, which equates the power's heating ability to a like amount of heating from a DC source. The trick is in making rms readings.

RF Ammeter Methods. The thermocouple RF ammeter (see Fig. 1) is an inherently rms-reading device because it relies on heating a very low value resistive heating element (R). The temperature of the heating element is measured by a thermocouple device (TC in Fig. 1), which produces a voltage proportional to the temperature of the thermocouple junction. A DC voltmeter is used



Your transmitter's RF power output is a key indicator of its health, and here are the circuits, instruments, and techniques you can use to accurately "take its temperature."

JOSEPH J. CARR

to measure the thermocouple output, but its scale is calibrated in units of current (amperes, milliamperes).

When the RF ammeter is in series with the transmission line from the transmitter to a resistive load (R_L) or a resonant antenna, the rms RF power level can be calculated from $I^2 \times R_L$. The advantage of the thermocouple RF ammeter is that it can be used with any load resistance, while other meters are designed for a specific load (usually 50-ohms). The disadvantages include the need to make a calcula-

tion, and the fact that RF ammeters tend to fade out above some frequency in the 40- to 70-MHz range.

Calorimeter/Bolometer Methods.

A number of professional-grade RF-power meters are based on the fact that the temperature change in a resistive load is proportional to the rms value of the applied RF waveform. Figure 2 shows a basic form of calorimeter or bolometer: A heat-dissipating resistor with a resistance value equal to the desired load impedance is enclosed in an assembly with some sort of temperature-measurement device.

Theoretically, you could put a big dummy load in a room and use a glass-mercury thermometer to measure the air temperature of the room before and after the power was turned on, but that's hardly practical. If, however, you embed a dummy load and some temperature sensor (thermistors and thermocouples are used) in a small assembly, then the before and after temperature rise of the resistor can be measured.

A low-cost instrument can be built using only the dummy load and Temperature Sensor 1, but that would ignore the problem that ambient temperature also affects the measurement sensor. It is not uncommon to include a second sensor to measure ambient temperature, so that changes in ambient temperature can be factored into the measurement. The results can then be displayed on an analog or digital meter.

Some calorimeter methods use two sensors (three could be used if ambient is accounted for) in a comparison measurement. A low-frequency (e.g. 60-Hz) AC power source is used to drive one sensor/resistor, while the RF power is used to drive the other. A differential

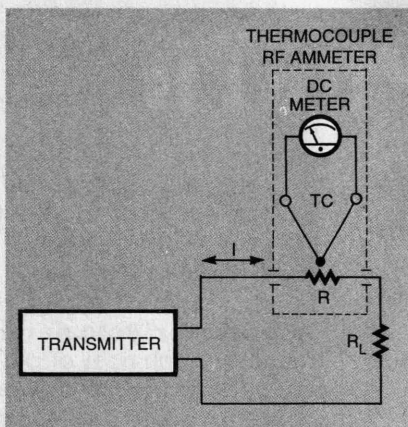


Fig. 1. The thermocouple RF ammeter relies on heating a very low value resistive heating element (R) to measure RF power.

meter will show when the two output levels are the same. At this point, the easily measured 60-Hz AC power level is equal to the applied RF power.

Simple Diode Detector Methods.

A simple diode detector scheme (see Fig. 3) can be used to measure the RF power applied to a load. The scheme shown was used on the Heathkit *Cantenna* dummy load that was popular some years ago, and is still used on similar products today.

The diode is an envelope detector, and produces a pulsating DC output from the RF voltage applied across the load (R_L). Capacitor $C1$ filters out the pulsations to pure DC. The power can be inferred from V_o^2/R_L .

The actual voltage applied to the diode is reduced by a resistive voltage divider ($R1/R2$) and is only a fraction of the applied voltage. This allows higher power levels to be measured. A diode such as a germanium 1N60, a silicon 1N914 or 1N4148, or a Schottky diode can be used for $D1$. Note that both $R1$ and $R2$ should be much larger than the load (R_L). Typical values for the circuit are $R1 = 100,000$ ohms, $R2 = 1000$ ohms, and $C1 = 0.05\text{-}\mu\text{F}$.

The diode detector circuit of Fig. 3 remains popular because it is simple and easy to implement. But it suffers from the fact that it measures the approximate peak power, not the rms power. On a sinewave CW signal, the rms power can be approximated by $(0.707 \times V_o^2)/R_L$.

52 In-Line RF Wattmeters. Perhaps

**TABLE 1—METER READINGS
BASED ON 100-WATT RMS CW CARRIER**

Waveform	PEV _{rms}	PEP(PEV ² / Z_o)	Thermal Power
CW	100/1.414 V	100 W	100 W
AM (100%)	200/1.414 V	400 W	150 W
AM (73%)	173/1.414 V	300 W	127 W
SSB (1 tone)	100/1.414 V	100 W	100 W
SSB (2 tone)	100/1.414 V	100 W	50 W

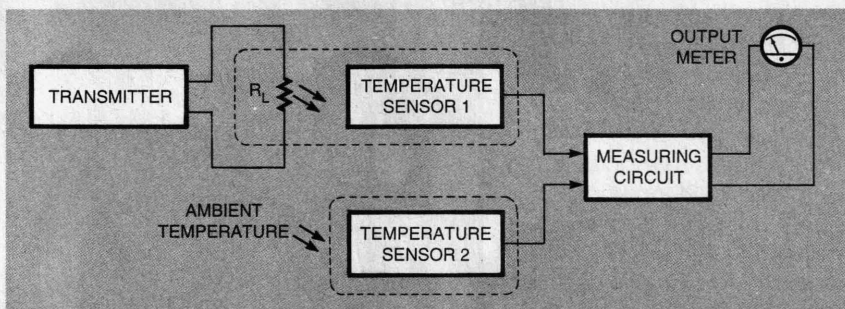


Fig. 2. In the bolometer/calorimeter approach to measuring RF power, a heat-dissipating resistor with a resistance value equal to the desired load impedance is enclosed in an assembly with some sort of temperature-measurement device. A number of commercial devices are based on this approach.

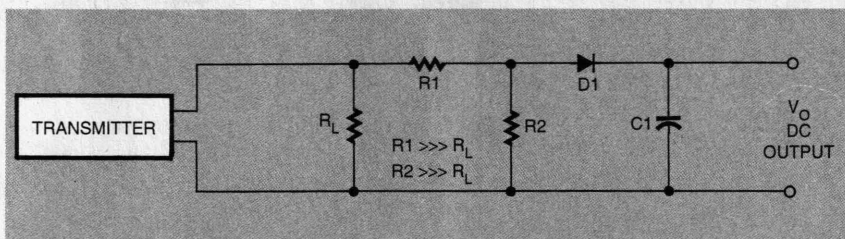


Fig. 3. A simple diode detector scheme shown here was used on the Heathkit *Cantenna* dummy load that was popular some years ago, and is still used on similar products today.

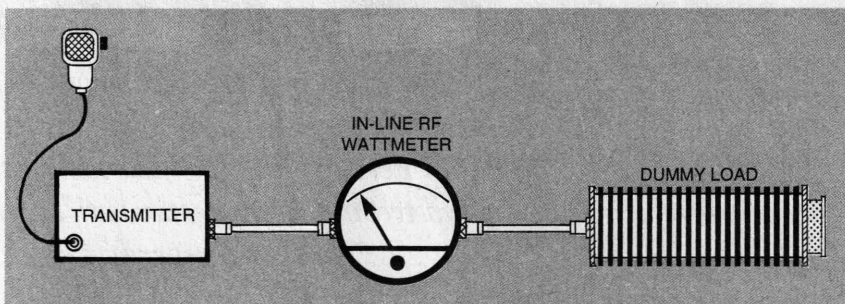


Fig. 4. Almost all RF wattmeters in use today are the in-line type.

the most common forms of RF power meter is the in-line wattmeter (see Fig. 4). The instrument is inserted in the coaxial line between the transmitter and either the antenna or a dummy load. Instruments designed for use with an antenna often have the ability to measure the forward and reflected power, so they can also be used to determine the standing wave ratio (SWR).

The classic Wheatstone bridge can theoretically be used for making an in-line RF wattmeter, but that is not a practical approach.

Such bridges are useful for making antenna impedance measurements at low power levels, but they cannot be left in-line because of the huge insertion loss involved. Other bridges, such as the micro-match bridge, discussed next, are used instead.

Micromatch Bridge. Figure 5 shows the basic capacitor-resistor micro-match-bridge circuit. Immediately we see that the micromatch is better for those applications than a conventional Wheatstone bridge because it only places a 1-ohm resistor

(R₁) in series with the transmission line. That resistor dissipates considerably less power than the resistors typically used in Wheatstone bridges. Because of that low-value resistance we can leave the micromatch in the line while transmitting.

As with Wheatstone bridges, the ratio of the resistances and/or reactances in the arms must be equal to create a null output to the meter. In that case, the ratio of the capacitive reactances of C₁ and C₂ must match the ratio of R₁ and the antenna or load resistance R_L:

$$X_{C1}/X_{C2} = R1/R_L$$

For a 50-ohm load, the ratio is 1/50, while for 75-ohm loads it is 1/75. A compromise situation that yields a small error on both 50-ohm and 75-ohm systems is to use a 68-ohm value for R_L and make the ratio $X_{C1}/X_{C2} = 1/68$. These ratios occur when C₂ ≈ 15 pF for 50-ohm systems, C₂ ≈ 10 pF for 75-ohm systems, or C₂ ≈ 12 pF for the compromise 68-ohm value.

The sensitivity control can be used to calibrate the meter. For fixed power meters, that potentiometer can be a trimmer type that is set when the meter is calibrated, and then left alone. At least one commercial micromatch bridge that was once popular with CB service technicians used a three-position switch to select three different sensitivity controls for 10-watt, 100-watt or 1000-watt power levels. Each range has its own sensitivity control, and those are used as needed.

Monomatch Bridge. The monomatch bridge of Fig. 6 is one of the "instruments of choice" for hams and other users for measuring RF power in the HF and low VHF ranges. In the monomatch design, the transmission line is segment "B", while the directional coupler transmission line segments are "A" and "C". The directional coupler lines are used for sampling the forward and reverse RF signals. Although some instruments used modified coaxial transmission lines, later versions use PC-board elements for A, B and C.

The sensor unit is basically a directional coupler with a diode detector element for both forward

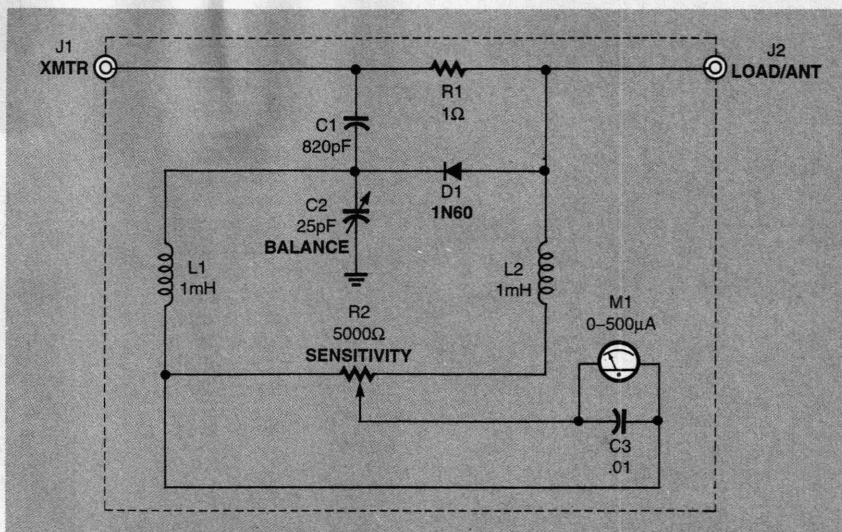


Fig. 5. The micromatch RF wattmeters offer important advantages over traditional Wheatstone bridges in RF power meter applications because they dissipate less power, allowing them to be left in the line.

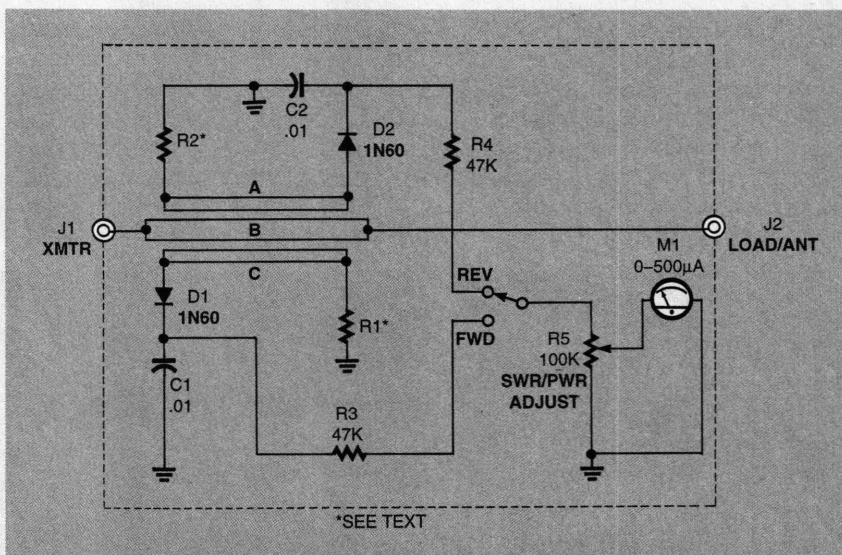


Fig. 6. The monomatch RF wattmeter is one of the "instruments of choice" for hams and other users for measuring RF power in the HF and low VHF ranges.

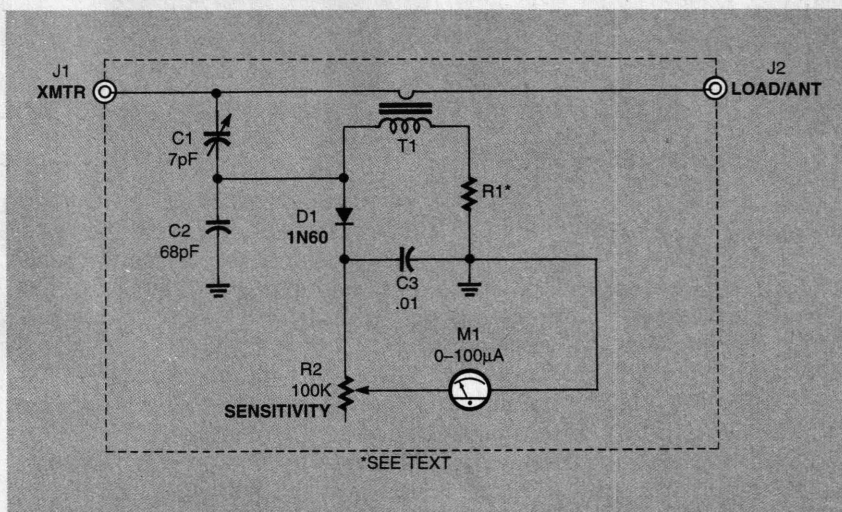


Fig. 7. In this version of the monomatch RF wattmeter, a single-turn ferrite or powdered-iron toroid transformer is used as the directional coupler.

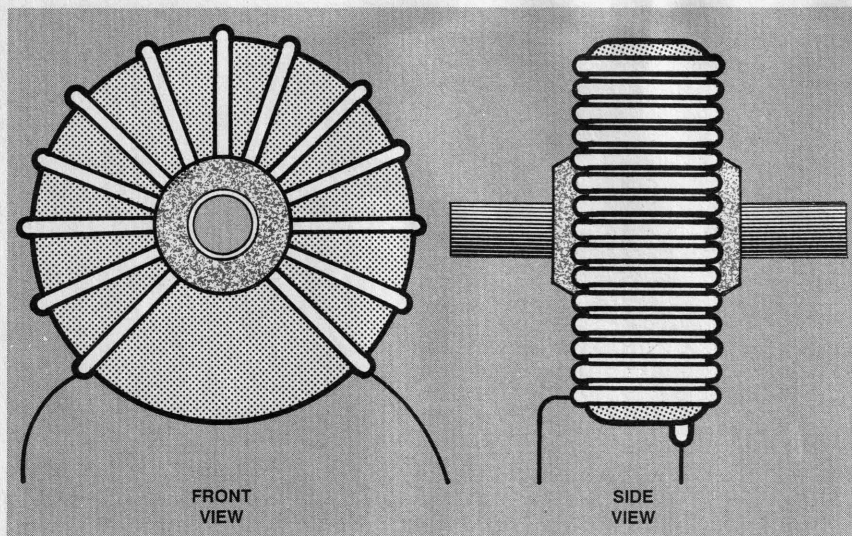


Fig. 8. Here are the winding details for the toroid transformer in the monomatch RF wattmeter in Fig. 7.

and reverse directions. For best accuracy, diodes D1 and D2 should be a matched pair, as should R1 and R2. The resistance of R1 and R2 should match the transmission-line characteristic impedance, although in many cases the 68-ohm compromise is used.

a transmission-line transformer as shown in Fig. 7. In that instrument, a single-turn ferrite or powdered-iron toroid transformer is used as the directional coupler. The transmission line passing through the hole in the toroid "doughnut" forms the primary winding of the transformer. The sec-

ondary winding is a single conductor passing through the hole. It is common to find $1/8$ -, $3/16$ -, or $1/4$ -inch brass tubing used for the primary. Note: When counting turns on a toroidal transformer, each pass through the hole is a "turn," so passing a straight wire or tube through the toroid hole once counts as one turn.

The value of R1 should match the transmission-line impedance, although, as usual, the 68-ohm compromise is often seen. If you opt to use the exact value in any of those circuits, then you can use either a single 51-ohm resistor, or two 100-ohm resistors in parallel. If you can find a precision 50-ohm resistor, however, use it (in standard carbon-composition or metal-film resistors, 51 ohms is a standard value, but 50 ohms is not).

Complex Waveforms. Measuring the RF power of unkeyed CW waveforms is relatively easy, but when modulation is applied, many instruments will read incorrectly. Table 1

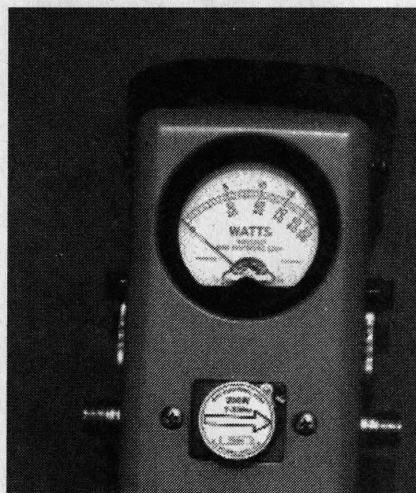


Fig. 9. The Bird Electronics Model 43 RF wattmeter is an old war horse that has served the author well over the years (photo courtesy of Bird Electronics Corporation).

The particular version shown in Fig. 6 uses a single DC meter movement to monitor RF power. With the addition of the switch and potentiometer R5, the circuit becomes both an SWR meter and a forward/reverse RF-power meter. Many modern instruments use two meter movements, one each for forward and reverse power.

A variation on the theme uses



Fig. 10. Bird Electronics Model 4421 RF wattmeter (photo courtesy of Bird Electronics Corporation).

ondary winding consists of 10 to 20 turns of small gauge enameled wire and is connected to a measurement bridge circuit (C1, C2, plus the load) that produces a diode-rectified output voltage.

Details for the construction of the sensor assembly is shown in Fig. 8. The secondary winding, made of 24- to 30-gauge enameled wire, is wound as shown in Fig. 8, with at least a 30 degree separation between the ends to minimize distributed capacitance. A rubber grommet is in-

serted into the hole of the toroid. The primary winding is a single conductor passing through the hole. It is common to find $1/8$ -, $3/16$ -, or $1/4$ -inch brass tubing used for the primary. Note: When counting turns on a toroidal transformer, each pass through the hole is a "turn," so passing a straight wire or tube through the toroid hole once counts as one turn.

Some Commercial Instruments.

Figure 9 shows a commercial RF power meter that the author bought nearly thirty years ago and has used heavily ever since. It is the Bird Electronics Corporation (30303 Aurora

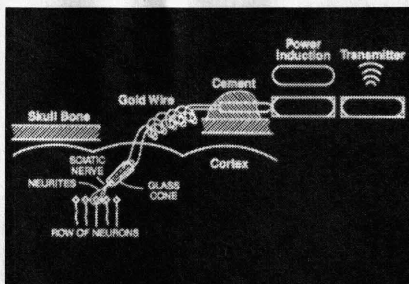
(Continued on page 56)

and developed using animals for over 12 years before it was implanted in humans. Both Federal Drug Administration and Institutional ethical approvals were obtained before the implants were made.

"The trick is teaching the patient to control the strength and pattern of the electric impulses being produced in the brain," Dr. Bakay said. "After some training, they are able to 'will' a cursor to move and then stop on a specific point on the computer screen."

First Tests. The electrodes have been successfully implanted into the brains of two patients at Emory University Hospital, one with amyotrophic lateral sclerosis (ALS or Lou Gehrig's disease) and one with a brainstem stroke. The first patient was able to control computer signals in an on/off manner for 76 days before she died from her terminal ALS condition.

The second patient, who is at the Atlanta Veterans Affairs Medical Center, is paralyzed except for his face due to a stroke and a subsequent heart attack. While dependent on a ventilator and unable to speak, he is fully alert and intelligent.



As shown in this block diagram, the electrode, which is housed in a cone-shaped glass enclosure, is implanted into the cortex of the brain and brain cells are induced to grow onto its tip. Neurons transmit an electronic signal when they "fire." Gold recording wires pick up the neural signals and transmit them through the skull. The system is powered by an induction coil placed over the scalp.

The electrodes were implanted in the part of the patient's brain that once controlled his arm and facial motion. All he has to do is think about moving his arm to create the electrical signals that are translated by the computer into movement of the cursor from icon to icon in a horizontal direction. As each icon is encountered, a phrase such as "See you later. Nice talking with you." is spoken by the computer.

The researchers are now working to help the patient communicate with the computer to produce

speech, as well as provide word processing, control of his environment, and maybe even access the internet. The future includes implanting the electrode in the brains of other "locked-in" patients.

Other Applications. The brain-implant technology has other applications. For instance, it could be used as a "spinal bypass" for those with spinal cord injuries that left them with stiff and uncontrollable muscles. Here neural signals could provide some control of electrical stimulators that activate the paralyzed muscles, therefore bypassing the area of spinal-cord injury.

The technique can also be used for basic research on understanding how the brain works. Never before have recordings been made from a human brain for so long and with such stability. For example, the firings of recorded neurons may change in response to differing inputs from different body parts: If neurons fire with taps to the arm, their firing may change over time from repeated taps to the face. This will help researchers understand how the brain communicates with other parts of the body. Ω

RF POWER

(continued from page 54)

Road, Cleveland, OH, 44139; Tel: 1-440-248-1200) *Model 43 Thruline* meter. It's a tough old war horse! That instrument is an in-line meter that uses a plug-in directional coupler and detector element to customize the meter for different RF power levels and frequency ranges. The tip of the plug-in element comes in close proximity to the coaxial line that runs inside the meter case from connector-to-connector. The analog meter displays the value of the RF power being measured.

The case of the *Model 43* has two sockets for additional plug-in elements to accommodate other power and frequency ranges. The heavy leather case that protects the meter in transit has space for several more elements. I have ele-

ments for amateur radio HF (10 watts for QRP, 100 watts for my transceiver, and 1000 watts for my linear amplifier), amateur radio VHF (50 watts), and lower power VHF (for marine radios).

Notice in Fig. 9 that the element has an arrow on it. The arrow points in the direction of power measurement. For example, if the load is connected to the right-hand coaxial connector, and the transmitter is connected to the left-hand coaxial connector, then the meter (as shown) will measure the forward power. If the direction of the arrow is reversed, then the reflected power can be measured. Once both the forward power (P_F) and reflected power (P_R) are known, the SWR can be calculated; but to make things easy, the *Model 43* comes with a nomograph in the instruction book that makes the calculation graphically.

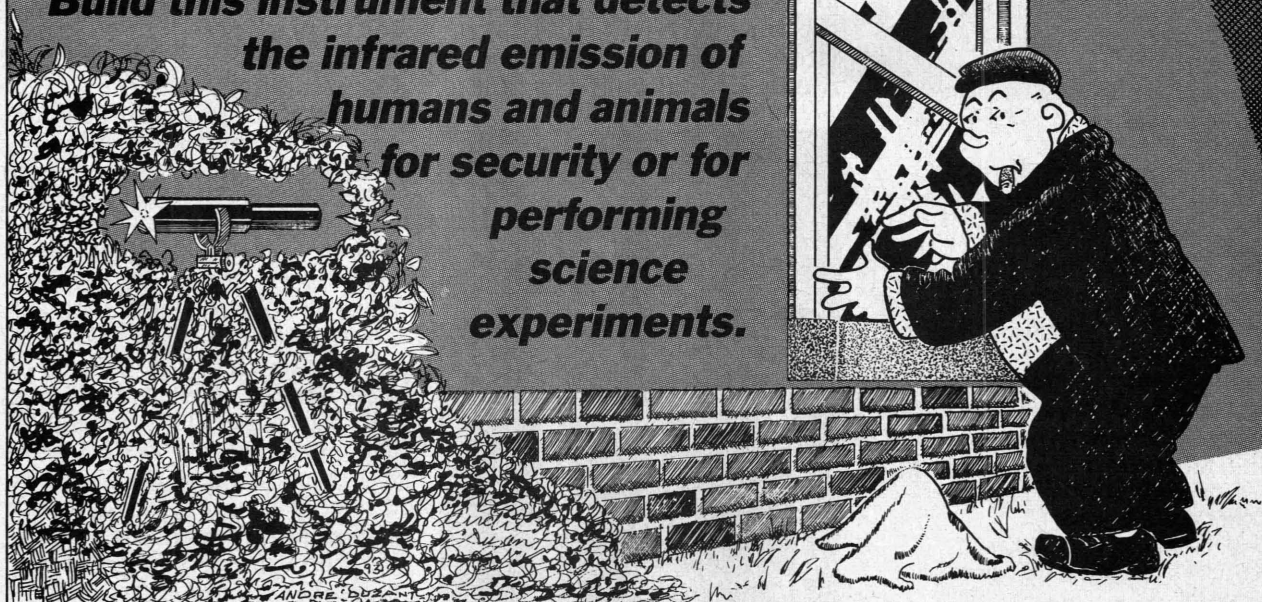
The Bird *APM-16 Advance Power Meter*, shown at the beginning of this article, is similar in concept to the older *Model 43*, but is considerably advanced. While the *Model 43* measures rms CW power, the *APM-16* will measure analog and digital complex waveforms, as well as CW (e.g. CDMA, TDMA, FDMA, COFDM and other modulation). It will measure both peak and rms power levels.

Figure 10 shows a Bird meter that uses a different approach. That meter uses a remote sensor head connected in-line between the source and load, and a multi-range digital readout display.

That wraps up our look at RF power-measurement circuits and instruments. Whether you need to measure RF power as part of your hobby or for your profession, keep the concepts we've discussed here in mind and you won't go wrong. Ω

Long-Range BODY-HEAT DETECTOR

Build this instrument that detects the infrared emission of humans and animals for security or for performing science experiments.



A PYROELECTRIC DETECTOR IN THIS battery-portable instrument detects the presence of animal or human intruders night or day by sensing their body heat. The detector can supplement your home security system, or you can monitor animals for nature study or find sources of heat loss.

This project will introduce you to the fundamentals of infrared emission and detection. The instrument could help you locate hidden or buried heat sources that could be wasting heat in your home. Detectors capable of "seeing" in the non-visible infrared region have proven themselves extremely useful in nighttime law enforcement and in military operations on land and in the air around the clock.

This instrument can detect the presence of humans or ani-

ROBERT IANNINI

mals up to several hundred feet away by sensing their body heat, which corresponds with a specific wavelength emission. Its narrow 8° field of view makes it suitable for detecting intruders passing through doors or moving along corridors inside a building. In addition, it can detect the presence of persons or animals entering the defined sensitive zone along roads or paths outdoors.

The best results will be obtained if the instrument is mounted on a camera tripod or other rigid support. When securely mounted, it can be panned manually or by electric drive over a wide sector to obtain a wider field of view. That makes it easier for the user to discriminate between true and false targets.

Pyroelectric detector

The heat sensitive elements in this instrument are two lithium tantalate (LiTaO_3) crystals within the TO-5 metal transistor case of the pyroelectric detector. The metal case has a rectangular silicon window in its cap for admitting infrared energy as well as a high-value resistor, and a low-noise field-effect transistor (FET).

The pyroelectric detector and pinout are shown in Fig. 1-a, and Fig. 1-b is a simplified schematic of the sensor circuit. Thermal compensation within the case prevents errors due to ambient temperature variations. The detector has a spectral range of 6 to 14 microns centered on 10 microns in the infrared band. This range is determined by the characteristics of the silicon window at the end of the TO-5 case.

PARTS LIST

All resistors are 1/4-watt, 10%, unless otherwise specified

R1—130 volts RMS varistor
R2, R3, R10–R12, R15–R18—10,000 ohms
R4—470,000 ohms
R5—18,000 ohms
R6—15,000 ohms
R7—270,000 ohms
R8, R9—2000 ohms
R13, R19—47,000 ohms
R14—10,000 ohms, trimmer potentiometer

Capacitors

C1, C2—0.2 μ F, 100 volts, Mylar
C3, C4—470 pF, 50 volts, ceramic disc
C5—0.22 μ F, 35 volts, tantalum
C6, C9, C10, C12—0.1 μ F, 50 volts, Mylar
C7, C8—15 pF, 50 volts, ceramic disc
C11—47 μ F, 35 volts, electrolytic

Semiconductors

BR1—1-amp, 200-PIV, full-wave bridge rectifier
D1—1N753, 6.2-volt Zener diode
IC1—PIC16C55, 8-bit microcontroller (Microchip)
IC2—MC145447, calling line identification receiver (Motorola)
Q1—MPSA13, NPN Darlington transistor

Other components

DISP1—16 $\times$ 1 LCD module
XTAL1—3.58-MHz TV colorburst crystal
S1–S3—SPST momentary push-button switch

Miscellaneous: Enclosure, PC board, IC sockets, 9-volt alkaline battery or 9-volt AC-to-DC wall adapter, telephone cord with modular plug, battery clip, ribbon cable, wire, solder, etc.

Note: The following items are available from Weeder Technologies, P.O. Box 421, Batavia, Ohio 45103:

- Double sided PC board (WTCID-B)—\$9.50
- Kit of all board-mounted components including pre-programmed PIC16C55 (WTCID-C)—\$24.50
- LCD display module (DISP16X1)—\$18.50

All orders must include an additional \$3.50 for shipping and handling. U.S. and Canadian orders only. Ohio residents must add 6% sales tax.

Note: Call your local telephone company to verify that Caller-ID service is available in your area before ordering.

data and changing the time from a 24-hour format to a 12-hour format, IC1 outputs the contents of its 16 registers to the LCD display through port C. The LCD display is wide enough to display only the time of day and the 7-digit telephone number, so the date and area code is discarded.

If the calling party is blocking the transmission of his number, a letter "P" will be received immediately after the date and time. In this case IC1 loads—along with the time—the word "PRIVATE" into its registers in place of a telephone number. Similarly, if the calling

returns the display to this block when a new call comes in. If IC1 detects an input from S3, it clears the display and erases all contents in the display memory.

The circuit can be powered by either a 9-volt battery or DC wall adapter. A regulator is formed by Q1, D1 and R19, which drops the 9 volts down to 5 volts. The regulator is specifically designed to draw low current for longer battery life. The circuit draws only about 1 milliamperes in the standby mode, and about 12 to 14 milliamperes when a call comes in.

Construction

The circuit is assembled on a 2.5- $\times$ 3-inch double-sided PC board. The foil patterns for the PC board are provided for those who wish to etch their own board. A pre-made PC board is also available from the source mentioned in the Parts List. Follow the parts-placement diagram shown in Fig. 4. Start by mounting IC sockets for IC1 and IC2. Pay particular attention to the pads on IC1 which have traces running between them. It is very easy to create a bridge of solder between those pads and the traces if you're not very careful.

Mount the resistors, capacitors and all other components. Pay attention to the orientation of the polarized components such as C5, C11, and D1. The pad spacing for Q1 is very tight, so extra care should be observed when soldering this component. Crystal XTAL1 should not be seated all the way down onto the board when soldering it because there is a chance that the metal case of the crystal will short the pads together. Use twisted pairs of insulated hook-up wire to connect the switches as shown. However, make sure the switches are installed in the case cover before you solder the leads to them.

Use a length of 14-conductor ribbon cable to connect the display module to the board. Separate the wires on each end of the cable for a length of about 1 inch to make it easier to solder them to the PC board and the display

Continued on page 84

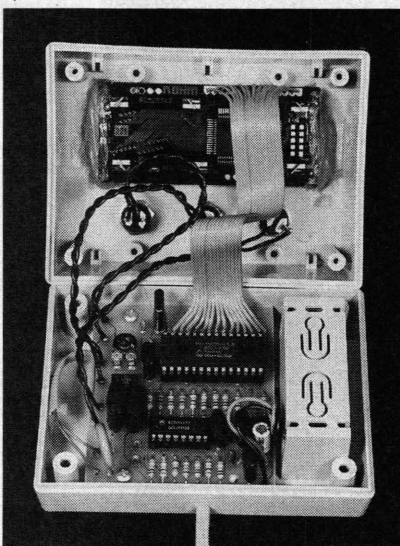
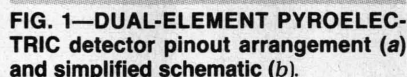


FIG. 5—THE COMPLETED PROTOTYPE. The small project case is the perfect size to place next to your telephone.

party is outside your area, a letter the letter "O" is received instead of the telephone number. IC1 then loads, along with the time, the message "OUT AREA" in its registers.

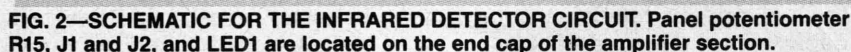
The LCD has its own internal memory which holds five 16-character messages. Each time a call comes in, IC1 addresses and writes to the next 16-character block, returning to the first block after reaching block number 5. Therefore DISP1 will always be holding the last five calls in memory. If IC1 detects an input from switch S1 or S2, it sends the instructions to DISP1 to shift the display backward or forward; it still remembers the address of the last block written to, however, and

To detect infrared emission, either the heat-emitting target



The infrared energy is focused on the sensitive rectangular window at the end of the detector case with a translucent plastic Fresnel lens, which has a focal length of approximately 5 centimeters. This translucent lens is transparent to infrared energy in the 10 micron region.

The thermal detector operates



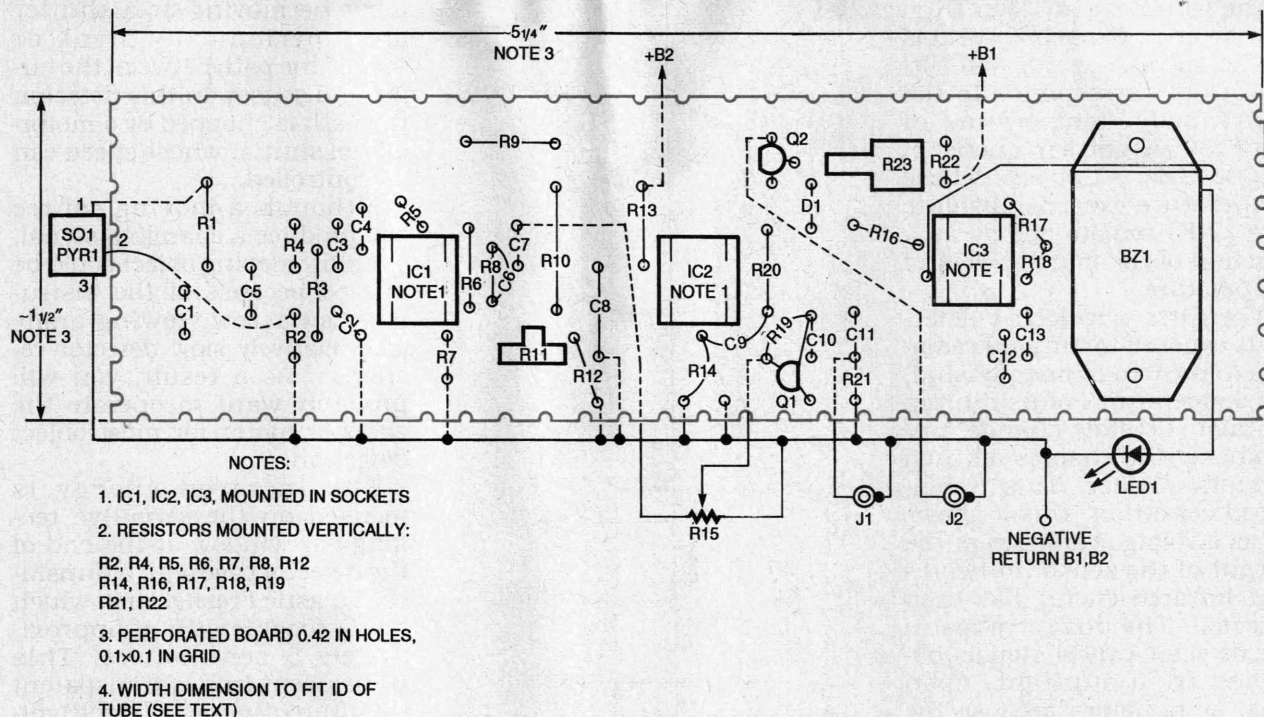


FIG. 3—LAYOUT OF THE CIRCUIT BOARD showing optimum component placement for point-to-point wiring. The circuit board width is determined by the inside diameter of the housing tube.

All resistors are 1/4-watt, 10% unless otherwise specified.

R1, R20—10,000 ohms
R2, R3, R5, R6, R7, R10, R12, R13, R14.
R16, R21—39,000 ohms
R4, R8—1,800,000 ohms
R9—1000 ohms
R11—25,000 ohms trimmer potentiometer, 8 mm vertical PC mount, Mouser Electronics 32RM403 or equivalent
R15/S1—1000 ohms control potentiometer, miniature panel mount, Mouser Electronics 31CT301 or equivalent
R17, R18—1,000,000 ohms
R19—390,000 ohms
R22—470 ohms
R23—200 ohms, trimmer potentiometer, 8 mm horizontal PC mount, Mouser Electronics 32RH202 or equivalent.

Capacitors

C1, C5, C8—1000 μ F, 16 volts, aluminum electrolytic
C2, C4, C7, C9, C11—4.7 μ F, 25 volts, aluminum electrolytic
C3, C6, C12—0.01 μ F ceramic disc
C10—0.047 μ F, 100 volts
C13—1 μ F, 50 volts

Semiconductors

PYR1—dual pyroelectric detector, 3-pin

TO-5 case, P2288 Hamamatsu Corp.
IC1—LM358N, dual operational amplifier, 8-pin DIP, National Semiconductor or equivalent

IC2—LM393N, dual comparator, 8-pin DIP, National Semiconductor or equivalent

IC3—NE555N timer, 8-pin DIP, Signetics or equivalent

Q1, Q2—N2222 NPN transistor

LED1—Light-emitting diode, T-1, yellow
D1—1N5230 Zener diode, 4.7 volts

Other components

BZ1—Piezoelectric buzzer, 6 volts, Mouser 25MS060 or equivalent

J1, J2—Jack, 3.5 mm ID

PL1, PL2—Plug, 3.5 mm OD

MOT1—DC motor, 0 to 12 volts, low-torque, slow speed, MCM Electric No. 58-500 or equivalent.

SO1—transistor socket, 3-pin (for PYR1)

Miscellaneous: Fresnel lens, polyethylene, 0.77 focal length, 1 1/2-inch, Fresnel Technical No. IR2 or equivalent, perforated circuit board, 0.1-inch grid (see text), plastic or metal tubing, end cap and inserts (see text), two sockets for 8-pin DIP ICs, T-1 LED

snap-in holder, 28 AWG insulated tinned copper wire (nine-strand ribbon cable (see text), two 9-volt battery clips with insulated wires, two 9-volt alkaline transistor batteries, tinned copper wire, plastic adhesive, plastic screw, solder.

Note: The pyroelectric detector is available from Hamamatsu Corp. Bridgewater, NJ (201) 231-0960. The Fresnel lens is available from Fresnel Technical, 101 Morningside Drive, Ft. Worth TX 76110, (817) 926-7474.

The following items are available from Information Unlimited, Box 716 Amherst, NH 03031 (603)-673-4730, fax 603 672-5406;
• A kit of all parts including tubing cut to length, circuit board, pyroelectric detector, motor and all active and passive components except batteries—\$69.50

• Pyroelectric detector and motor—\$12.50
Please add \$5.00 for shipping and handling. Allow 4 to 6 weeks for delivery.

in two detection modes: moving or stationary target. The moving detection mode permits the user to detect intrusions by moving people or animals. The detector's field of view in this

mode is narrow enough to be able to sense intrusions in restricted spaces, and its detection range is about 100 meters. In this mode it rejects constant background thermal energy.

The stationary mode permits the detection of stationary "hot spots" against a "cold" background such as a back yard or open space at night. Preferred in a search for hidden persons,

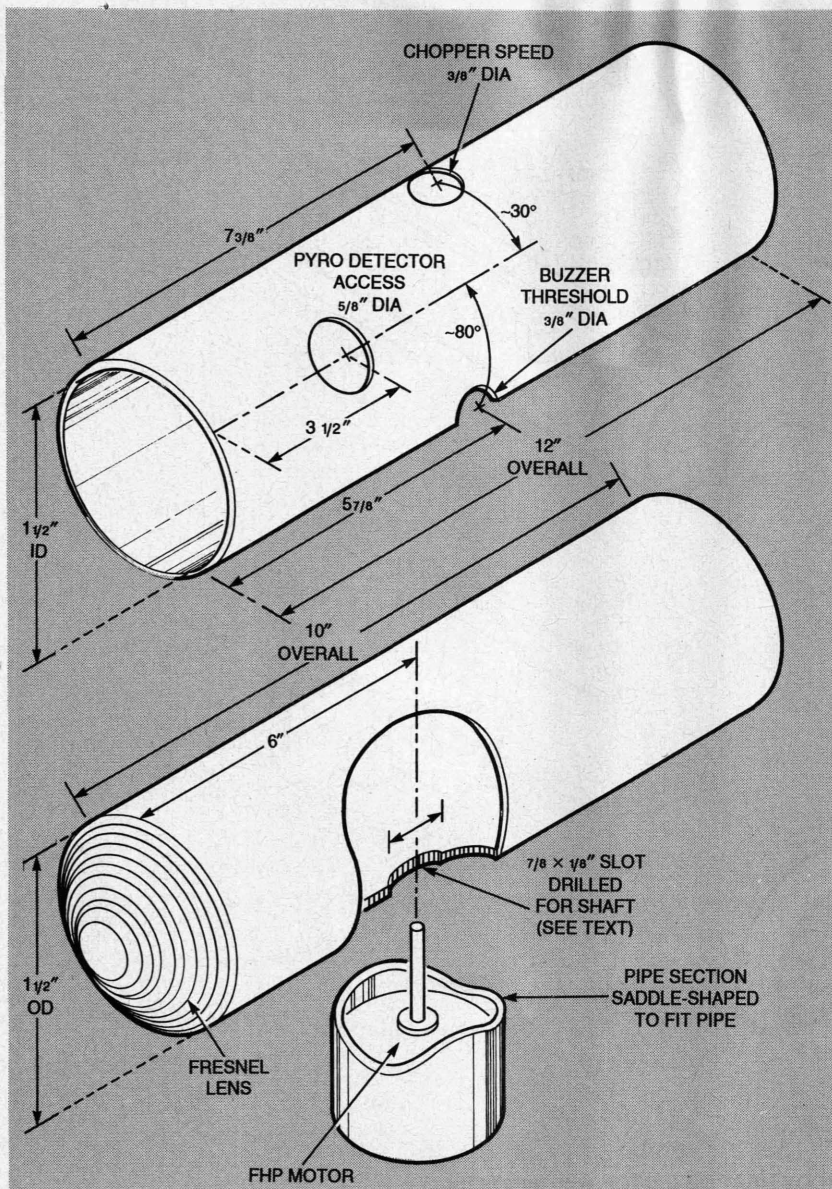


FIG. 4—CUTTING AND FORMING DIRECTIONS for the plastic or aluminum tubes for housing the sensor/amplifier and shutter/lens assemblies.

animals, or heat sources, this mode is so sensitive that some valid responses might seem to be false alarms.

System operation

Refer to the schematic, Fig. 2. The output of pyroelectric detector PYR1 is fed to the input of the low-frequency, dual-stage amplifier and filter IC1. The LM358N amplifier IC1-a, with a gain of about 2500, responds to a frequency of 0.1 to 10 Hz, peaking at about 1 Hz. This range matches the response of PYR1 to the infrared spectrum commonly generated by humans or large animals that is centered at about 10 microns.

The changing output of the filter IC1-b is further amplified by a factor of about 40 by PN2222 transistor Q1. This output can be capacitively coupled to an AC meter through EXTERNAL AC METER JACK J1. The output of J1 is an analog indication of signal strength, which should be between 20 and 100 millivolts.

The AC output of Q1 is also fed to IC2-a and b, an LM393N dual plus or minus "window" comparator. Trimmer potentiometer R11 sets the threshold activation level for buzzer BZ1. The output of IC2-a and b is sent to an AND gate, and its output is sent to the pin 2 TRIGGER

input of the NE555N timer IC3.

The output of IC3 can activate buzzer BZ1 and LED1 that are connected in series. LED1 illuminates when a target has been detected. Buzzer BZ1's on time is controlled by the 1-megohm timing resistors R17 and R18 across pins 8, 7 and 6, and capacitor C13 at pin 6. Panel potentiometer R15 (located on the end cap of the case) adjusts system temperature range and response to anticipated target size and temperature.

An optional chopper motor speed control consists of PN2222 transistor Q2 and trimmer potentiometer R23. Zener diode D1 provides a positive turn-on signal. Jack J2 provides DC drive for the shutter motor MOT1 through PL2.

External controls

Three access holes are formed in the tubular enclosure for the detector. One access port permits the lens position to be adjusted and the angle of PYR1 to be aligned for optimum results when the finished instrument is being set up. (This port is covered with an opaque band or tape to keep out ambient light when the instrument has been adjusted.)

Two other access holes are formed in the case to permit the two trimmer potentiometers R11 (buzzer threshold) and R23 (chopper motor speed adjust) to be set by a small insulated screwdriver or plastic trimmer adjusting tool. Panel-mount control potentiometer R15 can be set by a knob fitted on its shaft.

Chopper motor MOT1 speed can be set for optimum shutter chopping speed for detecting stationary objects by trimmer potentiometer R23. It produces the necessary "step function" in the infrared input signal.

Case material

Obtain a 12-inch length of aluminum or plastic tubing with an inner diameter of approximately 1.5 inches and a wall thickness of approximately 0.060 inch. Obtain a suitable plastic or aluminum cap with an inside diameter that will fit

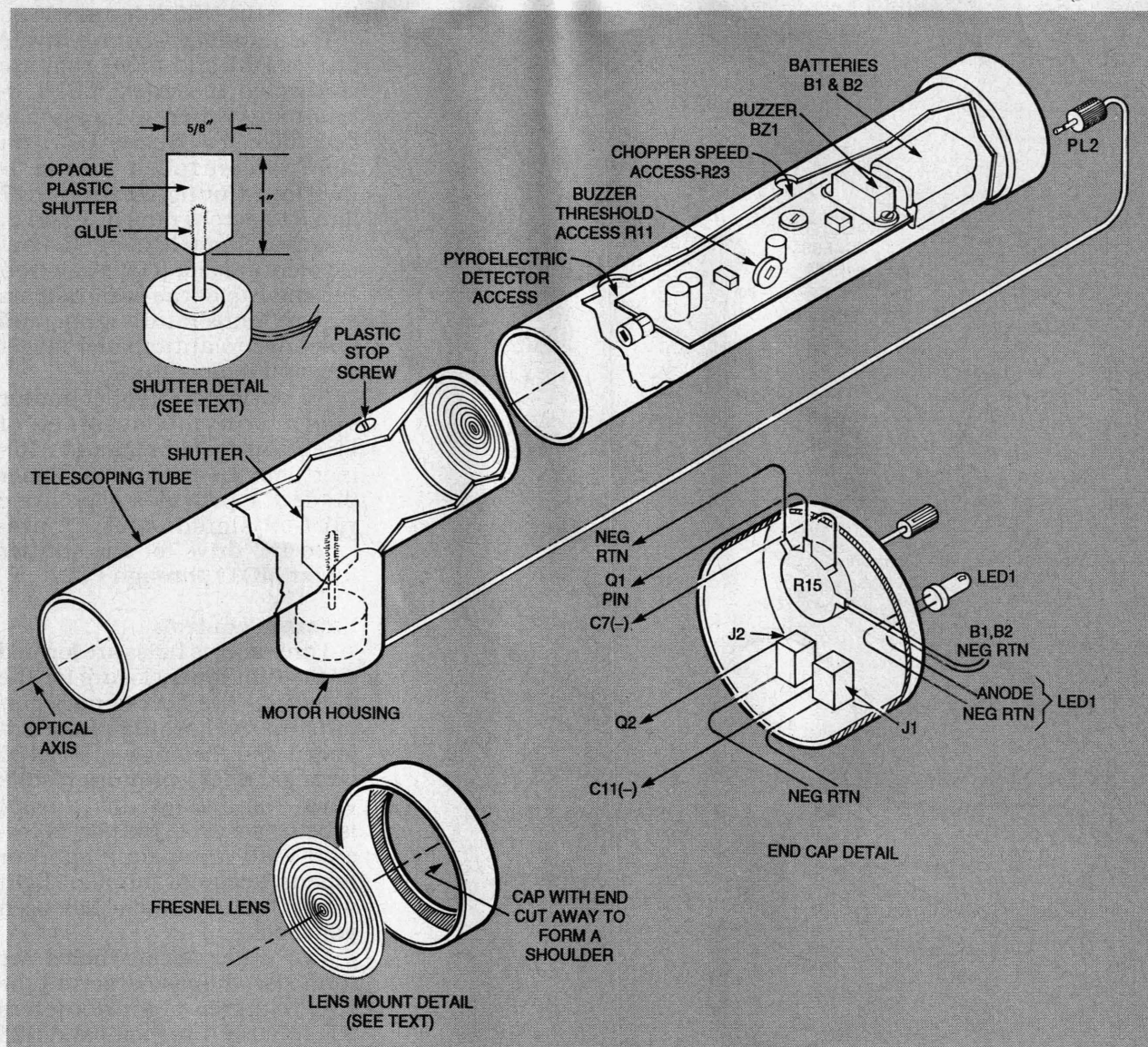


FIG. 5—CUTAWAY VIEW OF THE INFRARED BODY HEAT DETECTOR with details for making the shutter, lens mount, and end-cap assemblies.

snugly over the end of the tube.

Also obtain a 10-inch length of aluminum or plastic pipe with an exterior diameter of about 1.5 inches and a wall thickness of about 0.060 inch that telescopes snugly inside the larger tube. Then obtain an aluminum or plastic cap or cup about 1 inch deep that will fit snugly inside the smaller diameter tube.

Circuit board assembly

Refer to the electrical schematic Fig. 2 and the parts layout diagram Fig. 3. The amplifier circuit board was dimensioned to contain all of the electronic components and be able to slide inside the tubular case. The

prototype circuit board was cut from perforated board with a 0.1 inch grid to a length of 5¼ inches by 1½-inches, the approximate inside diameter of the housing tube. The width of the board should be cut slightly oversize. It can be sanded or ground down so that it can be press fit snugly inside the tube.

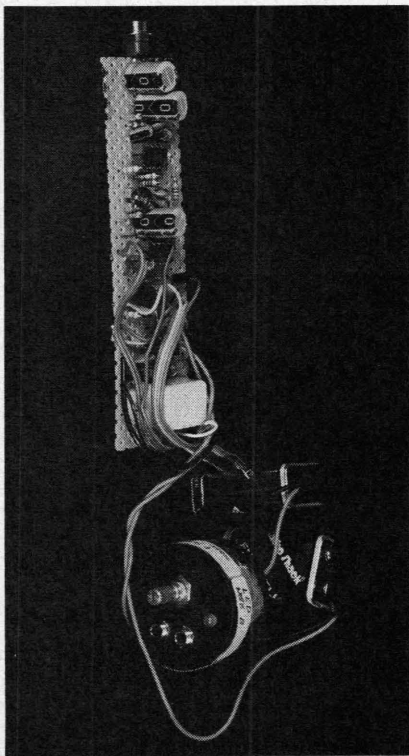
The parts layout as shown in Fig. 3 is intended to keep interconnection leads as short as possible. All connections are made with untrimmed component leads except for a bare wire bus that runs along the edge of the board on the wiring side. However, it might be necessary to solder tinned wire extensions on the ends of some sockets.

With Fig. 3 as your guide, insert the socket for pyroelectric detector PYR1 on the centerline of the board at the edge as shown. Then position the electronic components in the punched holes in the approximate positions shown in Fig. 3. In the prototype, resistors R2, R4, R6, R7, R8, R12, R14, R18, R19, R21 and R22 were all mounted vertically to conserve board space. One lead of each was bent back 180° to form a radial-leaded component.

Pay particular attention to the positions of the polarized devices and the pin 1 positions on the IC sockets. Bend the excess lead lengths on the wire side to form mechanical connections before doing any soldering.

Note the ground bus wire that connects pin 3 of SO1 with the negative sides of resistors R2, R7, R12, R14, R15, R21 and R23, as well as the negative sides of capacitors C2, C8, C12 and C13. In addition, this bus connects pin 4 of both IC1 and IC2 and the cathode of LED1 with one side of S1 (R15).

After completing all component insertion and soldering, check the circuitry visually against the schematic to be sure that all component placement is correct. Check for cold solder joints (dull gray color without evidence of solder flow), inadvertent solder bridges, or excessive solder on joints. Make all corrections before proceeding.



CIRCUIT BOARD AND END CAP assembly with batteries. The pyroelectric detector is at the top.

Mechanical assembly

Form the access holes in the 12-inch length of tubing as shown in Fig. 4. Then cut a slot that measures approximately $\frac{5}{8}$ -inch by $\frac{1}{8}$ -inch wide in the 10-inch tube as shown. Cut a small section of pipe that will accommodate the outside diameter of the DC shutter motor. Form a hole in the side of the section to accept the twin-wire

power cord, and shape the upper edge of the motor housing to form a saddle so that it will fit snugly against the pipe section.

Refer to the end cap detail in Fig. 5, and form the necessary holes in the tube cap to accommodate the two jacks J1 and J2, the light-emitting diode LED1 in a snap-in holder, and the combined potentiometer R15 and switch S1. Assemble those part in the cap and fasten them with the ring nuts provided.

End cap wiring

Cut approximate ten-inch lengths of No. 28 AWG insulated copper hookup wires required to connect the circuit board with the end cap components. The wires should be long enough to permit the end cap to be removed for replacing batteries B1 and B2 without removing the circuit board. In the prototype, nine-wire 28 AWG, multicolored flat-ribbon cable was used to make the nine connections from the circuit board to the cap.

Separate the individually color-coded insulated wires on both ends as necessary to make the appropriate connections. (If about an 8-inch length of flat cable remains bonded, it is easier to fold the cable back into the housing after final assembly.) Cut the black and red insulated wires from the 9-volt battery clips to about 10-inch lengths. Strip all wire ends, and twist all related pairs of wires before making the connections and soldering them.

Recheck the complete assembly looking for short circuits, cold solder joints, and improper location and orientation of polarized components. The circuitry can now be tested.

Electrical test

Install IC1 and IC2 in their sockets. Preset all potentiometers full counter clockwise with the pyroelectric sensor PYR1 out of its socket. Connect the 9-volt batteries B1 and B2 to their battery clips. Connect an ammeter on the 0 to 100-milliamperere scale in series with the common leads of both batteries across switch S1 at test point A (TP-A)

and measure the current. It should be between 15 and 20 milliamperes.

Turn the control knob of potentiometer R11 clockwise until the buzzer sounds. Expect to read a current of 60 milliamperes when LED1 is lit simultaneously. A voltage reading of 9 volts should be obtained at test point B (TP-B) and a voltage reading of 8.5 volts should be obtained at test point C (TP-C). Back off R11 slightly until the buzzer sound stops.

Insert pyroelectric detector PYR1 in its socket and slowly turn panel potentiometer R15 clockwise until the buzzer sounds. Keep PYR1 focused on a cold stationary background to prevent the detector circuit from responding erroneously to any movement. Make trial adjustments of R15 to verify the presence of a valid signal by passing your hand near the detector. Maximum sensitivity should be obtained when R15 is full clockwise and R11 is set to the critical activation point.

Connect a sensitive AC millivoltmeter set to the 500-millivolt range to jack J1, and watch for a response as an object or hand is placed in front of the detector.

Instrument housing

Refer to the lens-mount detail on Fig 5. Cut out the end of the aluminum or plastic cup that fits inside the 10-inch tube so that a shoulder about $\frac{1}{16}$ -inch wide is left as a retainer for the lens. Insert the lens in the aluminum sleeve and fasten it with one or two drops of a suitable adhesive. Then insert the lens assembly in the end of the tube as shown. Cement the assembly in place with an additional one or two drops of adhesive.

Refer to the shutter detail on Fig. 5. Assemble the chopper motor in the prepared housing. Then cut a thin piece of opaque plastic to the approximate dimensions shown, and glue it with suitable adhesive to the motor shaft as shown. After the adhesive has set, insert the shutter into the tube as shown in Fig. 5, and clamp the motor

Continued on page 87

HEAT DETECTOR

continued from page 43

housing assembly to the focus tube with plastic electrical tape. Then insert the focus tube in the larger circuit tube as shown in Fig. 5.

Operating instructions

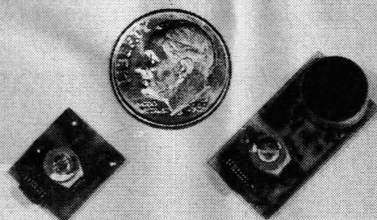
Be sure to allow the detector to "soak" in the ambient temperature in which it will be operated until it becomes temperature stabilized. For example, if it is to be used outdoors in the winter or summer when the temperature is significantly lower or higher than room temperature (22°C), allow the unit to remain in that environment for at least one hour and possibly as long as two hours before attempting to detect objects.

The ability of the detector system to discriminate a heat-producing stationary object from its background environment will depend on chopper-motor speed. Maximum sensitivity is achieved with one-tenth chopper revolution per second, but a chopper speed of 1 to 3 rps is recommended.

However, if you elect not to use the chopper for some application or experiment, look into the open end of the tube and move the shutter so that it is aligned in the focal plane to minimize interference with the incoming infrared emission.

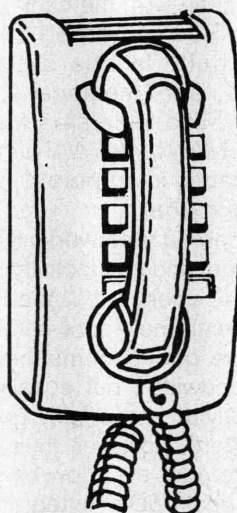
The field of view of this instrument with the Fresnel lens and pyroelectric detector specified is approximately 8°. Larger objects can be detected by panning the unit horizontally after mounting it on a rigid tripod.

After you are satisfied that the focus tube has been adjusted to the optimum focal length, mark the location of the end of the larger diameter tube on the smaller diameter tube. Drill a small hole in the end of the smaller diameter tube and insert a small plastic screw to act as a stop, as shown in Fig. 5. Either apply a patch of black plastic electrician's tape or slide an opaque plastic sleeve over the detector access port to keep out ambient light.



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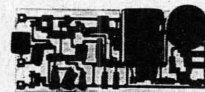


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RC FILTERS

**Understand
resistive-capacitive
filters and learn how to use
them in your projects
and experiments**

RAY MARSTON

THE TERM FILTER DESCRIBES A wide variety of frequency-selective circuits. Certain frequencies pass through a given filter while others are attenuated. There are four basic filter types: *low-pass*, *high-pass*, *bandpass*, and *band rejection* or *notch*. Filters composed of resistive (R), inductive (L), and capacitive (C) elements are called *passive filters*. *Active filters* include high-gain operational amplifiers with passive filter feedback networks.

Filter circuits that contain only resistors and capacitors are called resistive-capacitive (RC); those that contain only inductors and capacitors are called inductive-capacitive (LC). Filter circuits generally combine inductive and capacitive components because inductive reactance increases with frequency, and capacitive reactance decreases with frequency. The two opposing effects permit many possibilities in all filter design.

However, this article will focus on RC filters and applications. Later articles will review LC filters and look at active fil-

ters in greater detail. The big advantage of active filters is that they do not require inductors, which can be large and heavy at low frequencies. Active filters need only R and C components, but they require some kind of power supply.

A capacitor by itself has inherent filtering capability for alternating current because capacitive reactance, X_C , is inversely proportional to frequency

$$f_c = 1/2\pi RC$$

It blocks direct current completely and opposes the passage of low-frequency signals although signal passage becomes progressively easier as frequency increases.

Low-pass and high-pass

Filters contribute to the operation of many different circuits by screening out unwanted frequencies and allowing only the wanted ones to pass. Resistive-capacitive filters are better suited for low-frequency filtering (up to 100 kHz), whereas inductive-capacitive filters are better suited for high-frequency filtering (above 100 kHz).

Figure 1-a is the circuit of a

simple RC, low-pass filter that passes low-frequency signals but rejects those with higher frequencies. Resistor R1 is in series with the load, and capacitor C1, the reactive element, shunts the load. This filter exhibits a gradual rolloff beginning at the upper cutoff frequency where capacitive reactance equals the value of resistor R. Because it is a low-pass filter, there is only one cutoff frequency, and it can be determined by the formula:

$$f_c = 1/2\pi RC = 1/6.28RC = 0.159/RC$$

The cutoff frequency (f_c) is that frequency at which the signal output voltage is 6 decibels (dB) below its peak level.

Table 1 lists the formulas for determining f_c , R, and C for the schematics in this article that do not include component values. In these formulas 2π has been converted to the number 6.28.

The cutoff-frequency can also be measured at the *half-power points* as shown in Fig. 1-b. These are at 70.7% of the peak power with the real power dissipated at 50% of maximum. The

half-power point is the upper cutoff frequency of a low-pass filter.

High-pass filter

The high-pass filter passes high frequencies and opposes or blocks the passage of low frequencies. As shown in Fig. 2a, the simplest high-pass filter consists a single capacitor in series with the load and a resistor that shunts the load. Capacitor C1 opposes current flow that varies inversely with frequency. The higher the frequency, the smaller the opposition, measured in ohms. The filter completely or partially blocks signals at low frequencies, but permits their passage as frequency increases.

Figure 2-b shows the positive slope at the high-frequency end of the frequency vs. gain response curve for a high-pass filter. The *pass band* is defined as the area under the curve and the *stop band* is the area to the left of the curve.

The high-pass filter cuts off or blocks all frequencies below the cutoff frequency, f_c , permitting all those above that frequency to pass. The half-power (-3dB) point of a high-pass filter is the lower cutoff frequency. Both high-pass and low-pass filters have just one cutoff point, but as will be explained later, both *bandpass* and *band-reject* (or *notch*) filters have two cutoff frequencies.

Both of the filter circuits shown in Figs. 1-a and 2-a have a single RC stage and are known as a *first-order* filters. If a number (n) of these filters are cascaded, they will form what is known as an *nth-order* filter.

Phase-shift oscillator

Filters can be effectively cascaded by including them in the feedback networks of operational amplifiers. Figure 3 is a circuit for a third-order, high-pass filter that converts an op-amp into a phase-shift oscillator. The filter is inserted between the output and the input of the inverting (180° phase-shift) amplifier.

The filter will provide this phase shift at a frequency of

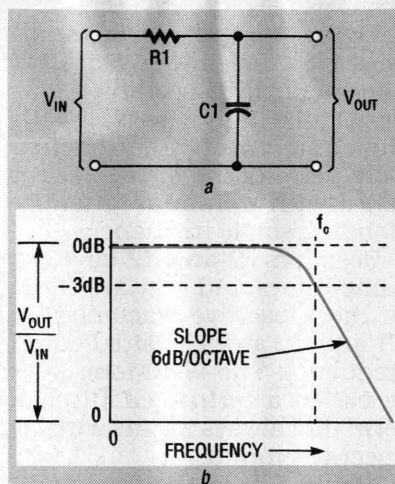


FIG. 1—LOW-PASS RC FILTER circuit (a), and frequency response curve (b).

TABLE 1
FORMULAS FOR DETERMINING
FILTER COMPONENT VALUES

High- and Low-Pass Filter		(Figs. 1 to 5)
Balanced Wien Tone Filter		(Figs. 6 to 10)
Twin-T Notch Filter		(Figs. 11, 14, and 15)

$f_c = \frac{1}{6.28 RC}$	}	$f_c = \text{kHz}$
$R = \frac{1}{6.28 f_c C}$		$R = \text{kilohms}$
$C = \frac{1}{6.28 f_c R}$		$C = \text{microfarads}$

about $1/4$ of the product of R and C. Thus the complete circuit has a loop shift of 360°, and it will oscillate at this frequency if the op-amp has sufficient gain. An op-amp with a gain of about $\times 29$ will compensate for filter losses and yield a loop gain greater than one.

Figure 4 is a schematic for an 800-Hz phase-shift oscillator. Potentiometer R4 must be adjusted to give a clean output sinewave and potentiometer R6 will vary the output gain.

Bandpass filter

A bandpass filter passes a specified frequency band while rejecting adjacent frequencies above and below that passband. A bandpass filter can be made by combining (or cascading) a high-pass filter with a low-pass filter as shown in Fig. 5-a so

that a band of frequencies not stopped by either filter is passed.

A typical bandpass filter response curve, as shown in Fig. 5-b, has a generally trapezoidal shape with a positive slope on the low-frequency (left) end indicating the limit of the high-pass stopband and a negative slope on the high-frequency (right) end defining the low-pass stopband. The flat top of the curve (0 dB) indicates constant signal gain.

The bandwidth of the filter is the frequency difference between the half-power, or -3dB points. These are the points

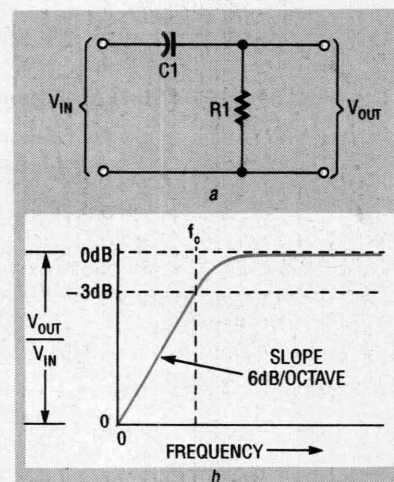


FIG. 2—HIGH-PASS RC FILTER circuit (a), and frequency response curve (b).

where the filter response is 3 dB down from the maximum point on the curve. The bandwidth is between f_{c1} (high pass and f_{c2}

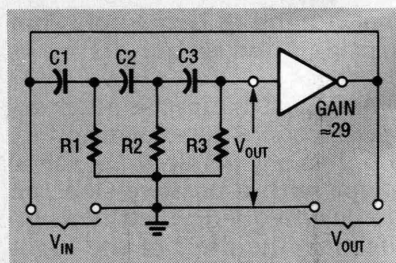


FIG. 3—THIRD-ORDER HIGH-PASS filter as a component in a phase-shift oscillator.

(low pass). The formula for bandwidth is:

$$\text{Bandwidth in herz} = f_{c2} - f_{c1}$$

The *passband* of a filter is the range of frequencies that pass through the filter. *Insertion loss* is the loss of signal strength experienced by the frequencies in the passband passing through the filter. If the filter were absent, and the source and load were connected directly to each other, the output signal would increase by the amount of the insertion loss.

Figure 6-a is a special band-pass filter called the Wien-tone filter made by cascading a low-pass and a high-pass filter with the same cutoff frequencies. This permits the filter to select tones with minimum attenuation at a single frequency.

Resistors R1 and R2 have the same values and they are equal to the capacitive reactances of C1 and C2 at the desired cutoff frequency. The Wien-tone filter is called a *balanced* filter, a term that always means with respect to ground.

Figure 6-b is the frequency-response curve of the balanced Wien-tone filter with an attenuation factor of 3 (-9.5 dB) at f_c . The circuit's principal feature is its ability to shift input signal phase $+90^\circ$ and -90° , and set it precisely at the f_c of 0° as shown in the phase-shift curve Fig. 6-c.

This filter can be combined with an operational amplifier to

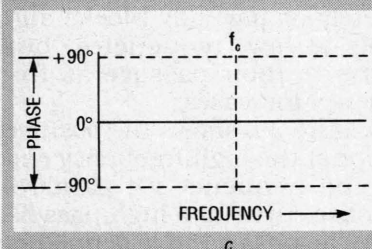
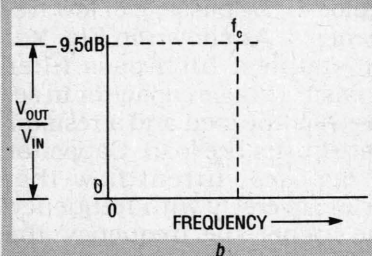
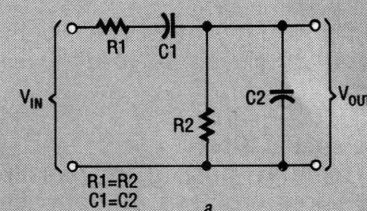


FIG. 6—BALANCED WIEN-TONE FILTER (a), frequency-response curve (b), and phase-shift curve (c).

become a sinewave oscillator as shown in Fig. 7. The output of the non-inverting amplifier with a gain of about $\times 3$ is fed back to its input through R1 and C1 to give a unity loop gain.

Band-reject filter

A *band-reject* or *notch* filter has a function that is the inverse of the passband filter. It is able to reject one specific frequency, the *stopband*, but pass all others. Figure 8-a shows a band-reject filter that is a modification of the circuit shown in Fig. 7.

In this circuit resistors R2 and R4 divide the voltage with a nominal attenuation factor of 3. As a result, the voltage divider and Wien filter outputs are identical at f_c , and the output, which equals the difference between the two signals, becomes zero.

The Wien-bridge band-reject filter network can close a loop around a high-gain operational amplifier to form an oscillator as shown in Fig. 9-a. It might appear that the Wien filter's output is fed to the input of the high-gain amplifier that has its out-

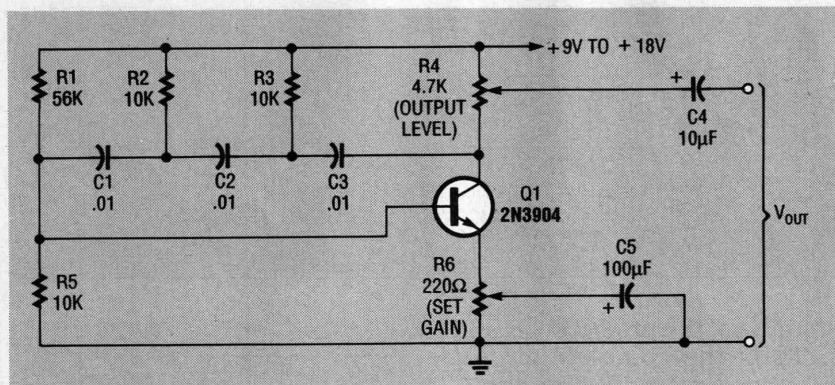


FIG. 4—PHASE-SHIFT OSCILLATOR produces 800Hz sinewaves.

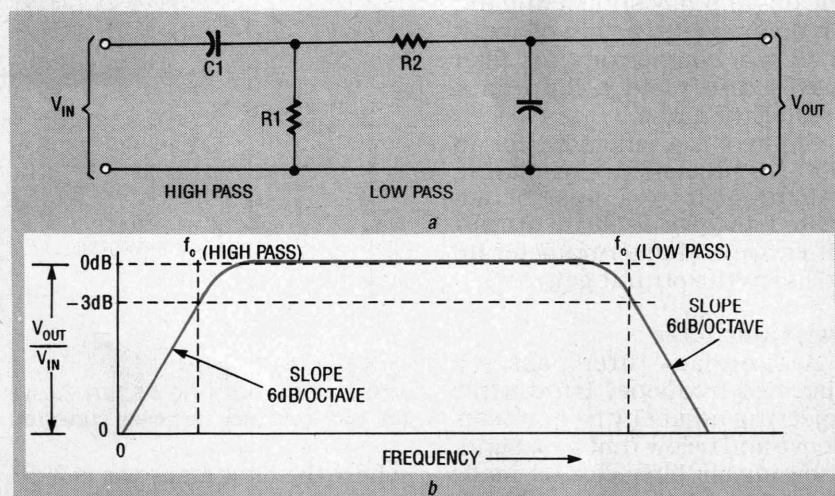


FIG. 5—HIGH-PASS AND LOW-PASS FILTERS cascaded to make a bandpass filter (a), and frequency response curve (b).

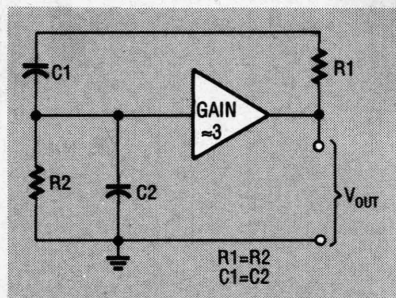


FIG. 7—BASIC WIEN-FILTER-BASED oscillator.

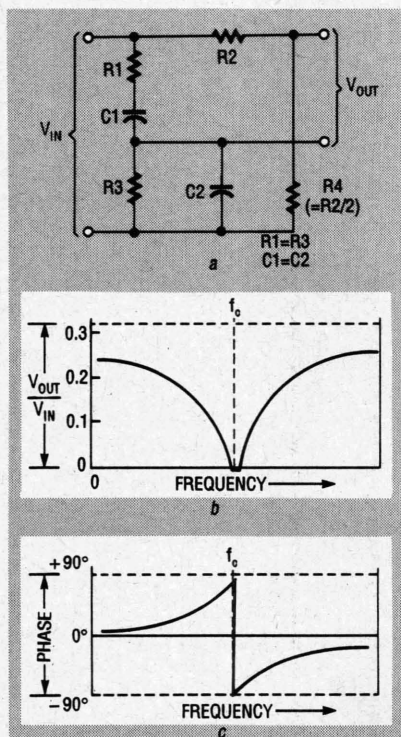


FIG. 8—WIEN-BRIDGE NOTCH FILTER (a), frequency-response curve (b), and phase-shift curve.

put fed back to the Wien filter's input to complete a positive feedback loop.

However, if the circuit is redrawn as in Fig. 9-b, it is clear that the op-amp actually functions as a $\times 3$ non-inverting amplifier, and that the circuit is similar to that shown in Fig. 7. Some form of automatic gain control will be needed if high-quality sinewaves are to be generated from this circuit.

The tuned frequency of the Wien-bridge network can be changed by simultaneously altering its two resistor or capacitor values. Figure 10 is the schematic for a wideband (15 Hz to 15 kHz) variable-notch filter. Ganged potentiometer R3 and

ganged switch S1 permit fine tuning and decade switching and trimmer potentiometer R6 performs null trimming.

Twin-T filter

Figure 11-a is a schematic for a twin-T notch filter. A prime advantage of the filter is that its input and output signals share a common-ground connection, and its off-frequency attenuation is less than that of the Wien-bridge filter. However, it also has a drawback: To work effectively, the values of all three resistors (and the capacitive reactances at a specified frequency) must be varied simulta-

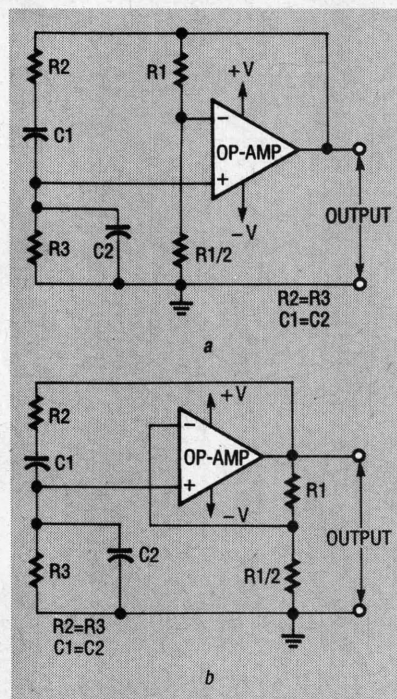


FIG. 9—THE WIEN-BRIDGE OSCILLATOR (a) is the equivalent of the oscillator shown at (b).

neously. This filter is *balanced* when its components have the precise ratios shown in Fig. 11-a. For perfect nulling, the $R1/2$ resistor value must be carefully adjusted.

The distinctive notched frequency-response curve for the balanced twin-T filter is shown in Fig. 11-b. Notice that at f_c the notch has a value of zero. The abrupt phase shift curve is shown in Fig. 11-c. The filter has zero phase shift at f_c , but that phase shift changes sharply to $+90^\circ$ or -90° for slight varia-

tions above and below f_c .

Both the balanced twin-T and the Wien-bridge filter have very low effective Q 's. The value Q is calculated by dividing the value of f_c by the bandwidth between the filter's two -3 dB points. A typical value for a twin-T filter is 0.24. The filter attenuates the second harmonic of f_c , 9 dB. By contrast, an ideal notch filter would not attenuate the input signal.

This shortcoming of the twin-T filter can be overcome by "bootstrapping" the common terminal of the filter, as shown in simplified block diagram Fig. 12. High effective Q values can be obtained with this circuit, and attenuation of the second harmonic of f_c will be negligible.

An explanation of how a balanced twin-T filter works can be quite complicated, so the equivalent diagram Fig. 13 is presented here to simplify that explanation. The filter has been resolved into a parallel-driven, low-pass filter ($f_c/2$) and a high-pass filter ($2f_c$) whose outputs are connected to an RC voltage divider (f_c). This output divider loads the two filters and affects their phase shifts.

As a result, the signals at points A and B are identical in amplitude, but have phase shifts of -45° and $+45^\circ$ respec-

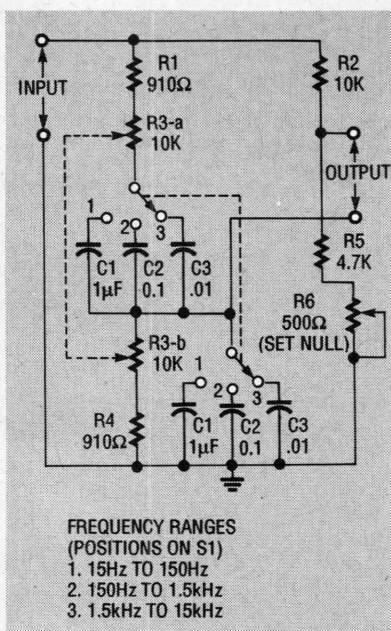


FIG. 10—VARIABLE-FREQUENCY, Wien-bridge notch filter (15 Hz to 15 kHz).

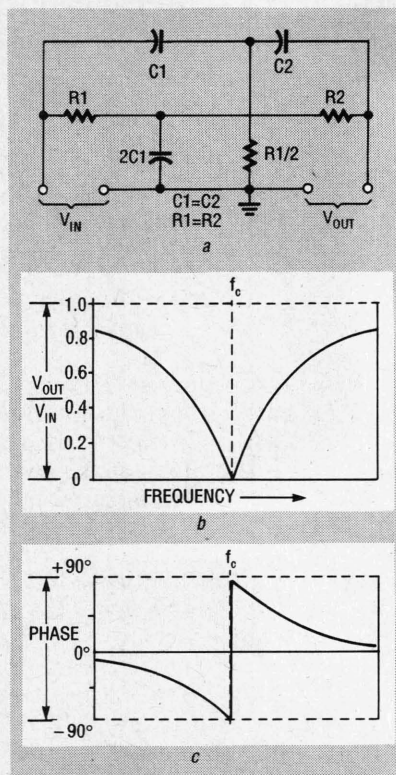


FIG. 11—BALANCED TWIN-T NOTCH FILTER (a), frequency-response curve (b), and phase shift curve (c).

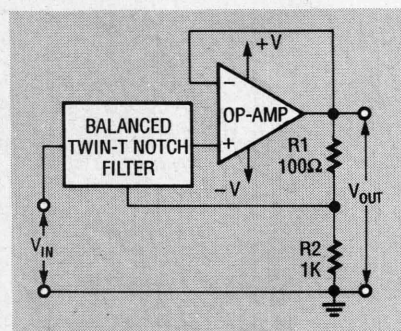


FIG. 12—BOOTSTRAPPED HIGH-Q notch filter.

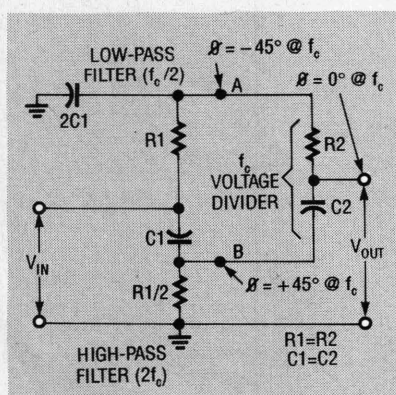


FIG. 13—EQUIVALENT CIRCUIT DIAGRAM of the balanced twin-T filter.

tively at f_c . At the same time, the impedances of the R and C sections of the output divider are equal and introduce a 45° phase shift at f_c . Thus the divider effectively cancels the two phase differences and gives a precise output of zero, the phase-cancelled difference in amplitudes between the two signals.

Therefore, a perfectly bal-

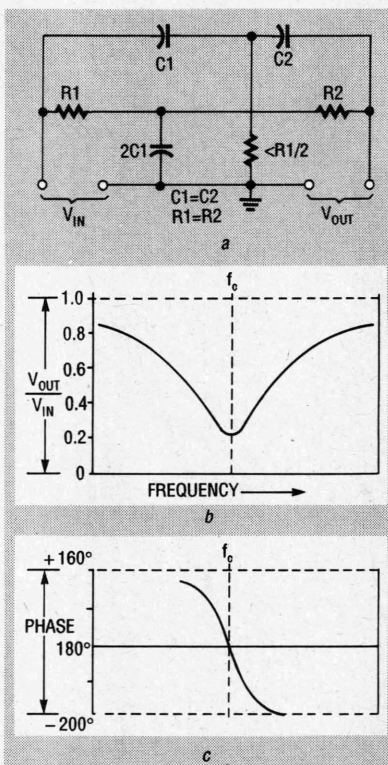


FIG. 14—NEGATIVELY UNBALANCED TWIN-T FILTER provides a 180° phase shift at f_c (a), its frequency-response curve (b), and phase-shift diagram (c).

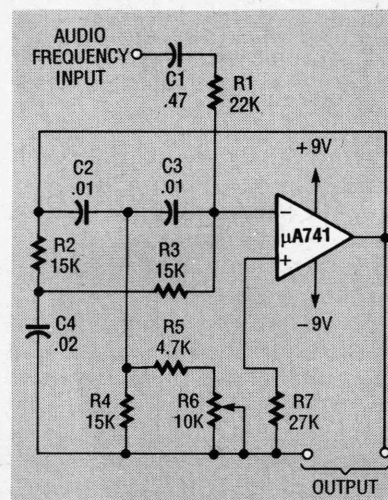


FIG. 15—A 1 KILOHERTZ OSCILLATOR/ACCEPTANCE FILTER based on a negatively unbalanced twin-T network.

anced twin-T filter has a zero output and zero phase shift at f_c . As shown in Fig. 11-c, if the frequency is slightly below f_c , the output is dominated by the response of its low-pass filter, and is shifted in phase by -90° ; at frequencies above f_c , the output is dominated by the response of its high-pass filter, and is shifted in phase by $+90^\circ$.

An unbalanced version of the twin-T filter can be made by changing the value of the $R1/2$ resistor to one that is not ideal. If this resistor has a value greater than $R1/2$, the circuit will be *positively unbalanced*. It will act the same way as was just described, but its notch will have limited depth (it will not reach zero). However, it still offers a zero phase shift at f_c .

If, by contrast, the resistor has a value less than $R1/2$, the circuit, as shown in Fig. 14-a, will be *negatively unbalanced*. It will also give a notch of limited depth, as shown in Fig. 14-b, but it has the useful characteristic of being able to produce a phase-inverted output. There will be a 180° phase shift at f_c , as shown in Fig. 14-c.

Figure 15 is a schematic of a negatively unbalanced twin-T notch filter that can be a component in either a 1 kHz oscillator or a tuned acceptance filter. The twin-T filter is connected between the input and output of the high-gain inverting amplifier so that an overall shift of 360° occurs at f_c .

The circuit will oscillate if potentiometer $R6$ is adjusted so that the twin-T notch gives enough output so the system has an overall gain greater than unity.

To convert the circuit to a tone filter, adjust trimmer potentiometer $R6$ to give a loop gain that is less than unity, and feed an audio input signal through $C1$ and $R1$. Under this condition, $R1$ and the twin-T filter interact to form a frequency-sensitive circuit that gives high negative feedback and low gain to all frequencies except f_c . The circuit gives little negative feedback and high gain at f_c . Trimmer potentiometer $R6$ can vary the sharpness of the tuning. Ω

Attention, Weather Watchers!

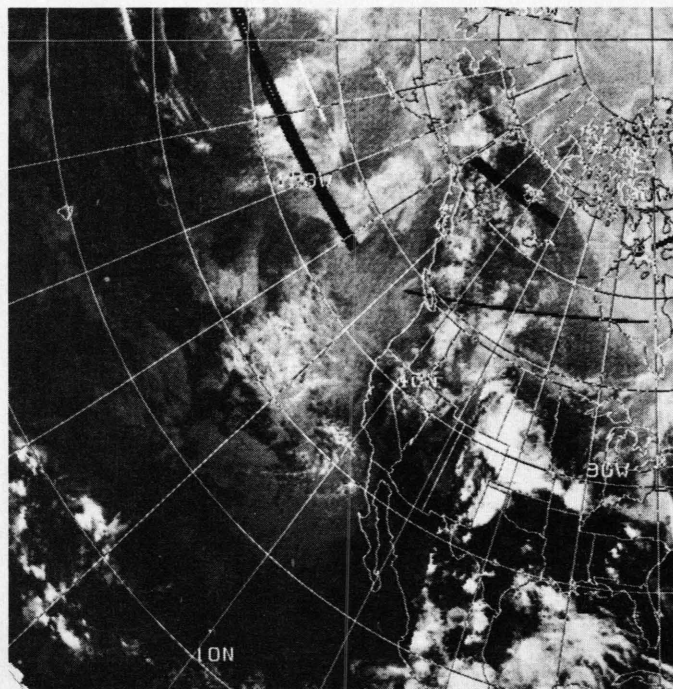
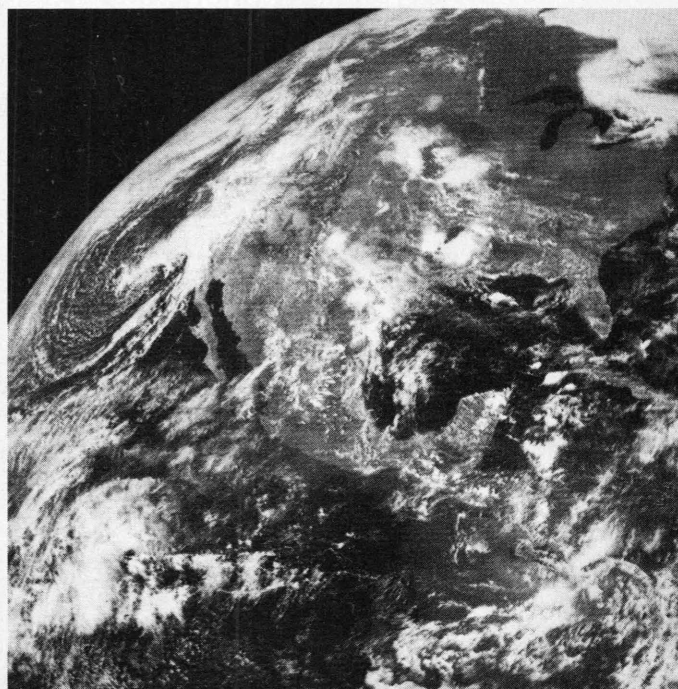
*—advanced circuitry
for WEFAX processing*

The purpose of this article is to present some solid state circuitry ideas on reproducing satellite APT/WEFAX pictures. I have felt for some time

now that there has been room for improvement in equipment currently in use. Most people have been resorting to a small army of vacuum tubes to

do the work in facsimile machines. Almost everyone is using a sensitive solid state receiver, so now it's time to improve the video processing. The

schematics on the following pages are currently being used here to process GOES WEFAX. They have been doing so since GOES WEFAX was initiated in the



summer of 1975 through SMS-1 on a regular schedule. Some of the features these circuits offer are:

1. Solid state design.
2. Closed-loop automatic gain control (agc).
3. Selectable agc time constants.
4. Precision video demodulator.
5. Low-ripple video filter with high out-of-passband attenuation.
6. High reliability.

These circuits by no means approach the state of the art; however, they can make a vast improvement in your satellite photos. The satellite photos displayed in this article were transmitted over the GOES WEFAX data link and were processed by my homemade facsimile machine. I feel that, with careful construction and planning, most homemade

facsimile machines can equal or surpass many commercial APT displays. The following discussion on video processing and the schematics should throw some light on what I am currently doing with GOES WEFAX.

Theory of Operation

In Fig. 1, you see the input video amplifier. The receiver input, tape-recorder input, and test-pattern generator switching relays are also shown.

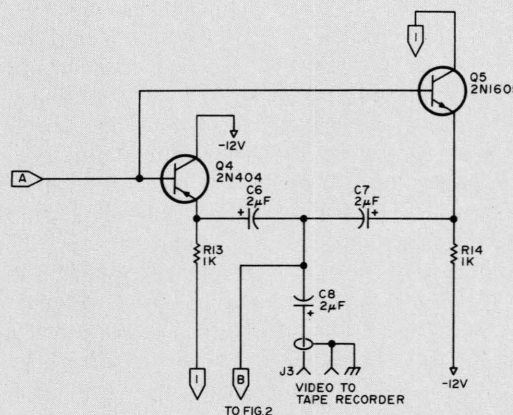
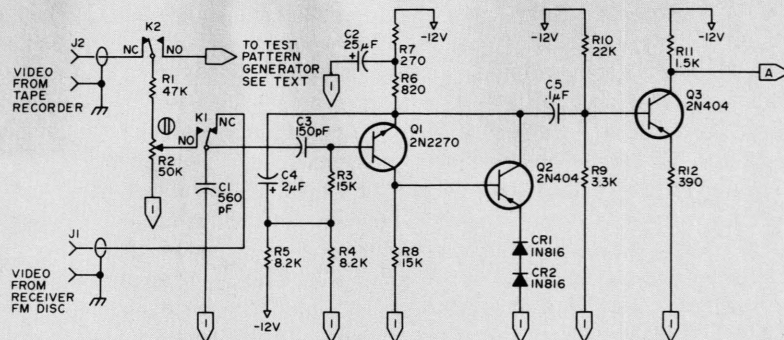


Fig. 1. Input amplifier and relay switching.

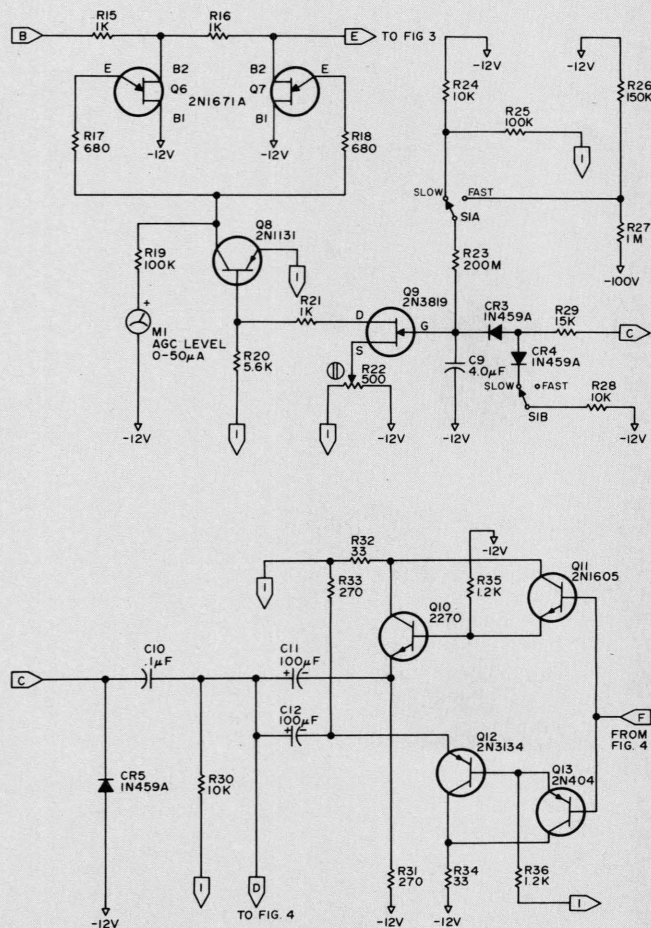


Fig. 2. Electronic attenuator and agc detector.

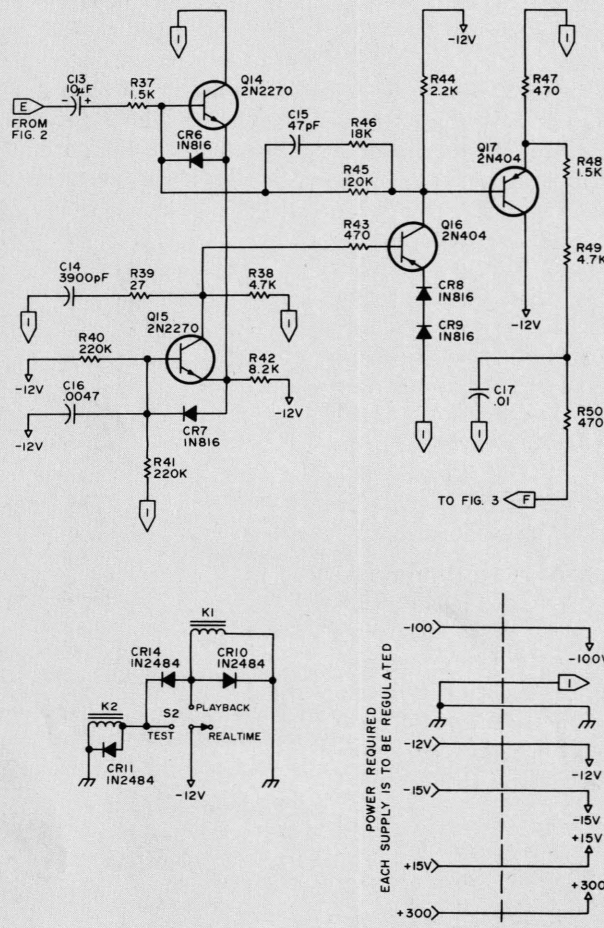


Fig. 3. Agc amplifier and mode selector.

Switch S2, in Fig. 3, selects the operating mode, or the position K1 or K2 is in. For now, assume S2 is in the real-time mode. With S2 in real time, K1 and K2 are normally closed. Video is now supplied from the FM discriminator in the receiver to J1 and into the video preamplifier. Transistors Q1, Q2, Q3, Q4, and Q5 amplify the input signal from the receiver sufficiently to drive the J3 output. Output J3 supplies amplified video to a tape recorder. The high-impedance line input on the tape deck should be connected here for recording. Transistors Q1 and Q2 form a Darlington pair; Q4 and Q5 are connected in a complementing circuit. Output J3 is driven by Q4 and Q5 and coupled to the signal via capacitors C6, C7, and C8. The negative end of C8 is supplied to Fig. 2, point B. I will pick up here in a moment with Fig. 2, but, first, I still have two more operating modes on S2 to discuss.

With S2 in real time, my station is acquiring spacecraft data and recording it on tape. In order to play back previously recorded data, S2 is switched to playback. Relay K1 now closes, and the satellite video is played back into J2 from the tape-deck line output. Pot R2 is used to limit the played back signal to the same approximate level as that at which the receiver output is during real-time operation.

The final position on S2 is a test mode. This input is connected to the output of my grey-scale bar generator. It provides the machine with a proper test signal for calibration purposes (see 73, Apr., '78).

Point B on Fig. 1, as stated earlier, connects to point B on Fig. 2. This is the input to the electronic attenuator. The attenuator output, point E, drives a

high-gain agc amplifier, shown in Fig. 3. Its open-loop gain is about 80. The amplifier output, point F, feeds a four-transistor power amplifier consisting of Q10, Q11, Q12, and Q13 on Fig. 2. This amplifier provides drive to the agc peak detector, CR3. The peak detector is used to charge C9 to the peak amplitude of the video signal at point D in Fig. 2. The charge on C9 is slowly bled off by R23. Transistors Q9 and Q8 amplify the value of the charge and produce the voltage/current source needed to control the UJT attenuator. The discharge path of C9 is controlled by S1. Fast decay is for GOES WEFAX, while slow decay is for NOAA radiometer data.

Now, to simplify confusion somewhat, I will try to sum up the purpose of the agc circuit. The main idea here is to control the peak-to-peak level at point D. The agc circuit, when properly adjusted by R22, will hold an unmodulated 2400 Hz input at 3.3 volts p-p on point D. When video variations are present, capacitor C23 does not have time to

charge to 3.3 volts p-p because of the controlled decay or discharge path. The video sync, however, will be at 3.3 volts p-p, since it occupies a substantial period of time. The agc circuit is used to compensate for large changes of video peak-to-peak output from a receiver. Also, it sets up a maximum and

minimum signal level with which to work. In this case, signals fall between 3.3 volts p-p and the system noise floor.

Knowing these levels, the video detector, filter, and lamp driver were designed. The agc output at point D drives the precision video demodulator on Fig. 4. The incoming

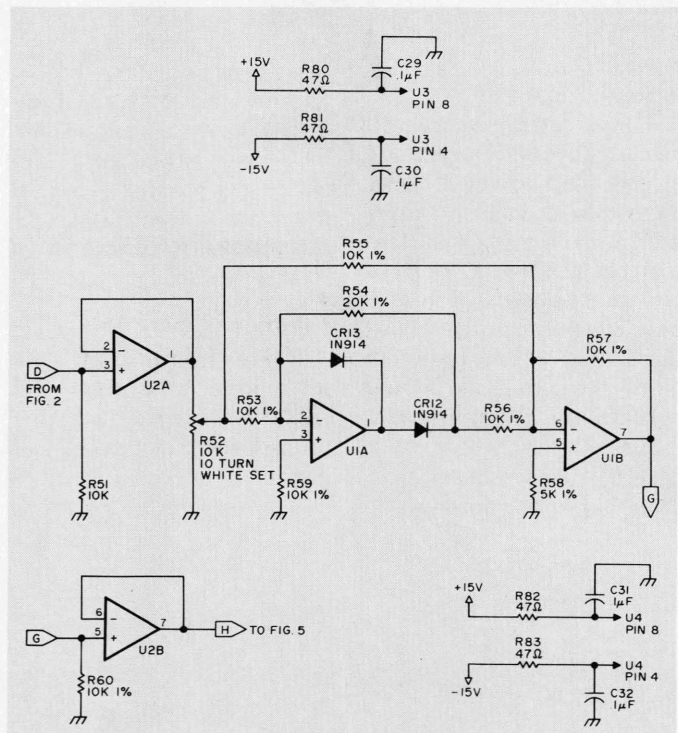


Fig. 4. Video demodulator.

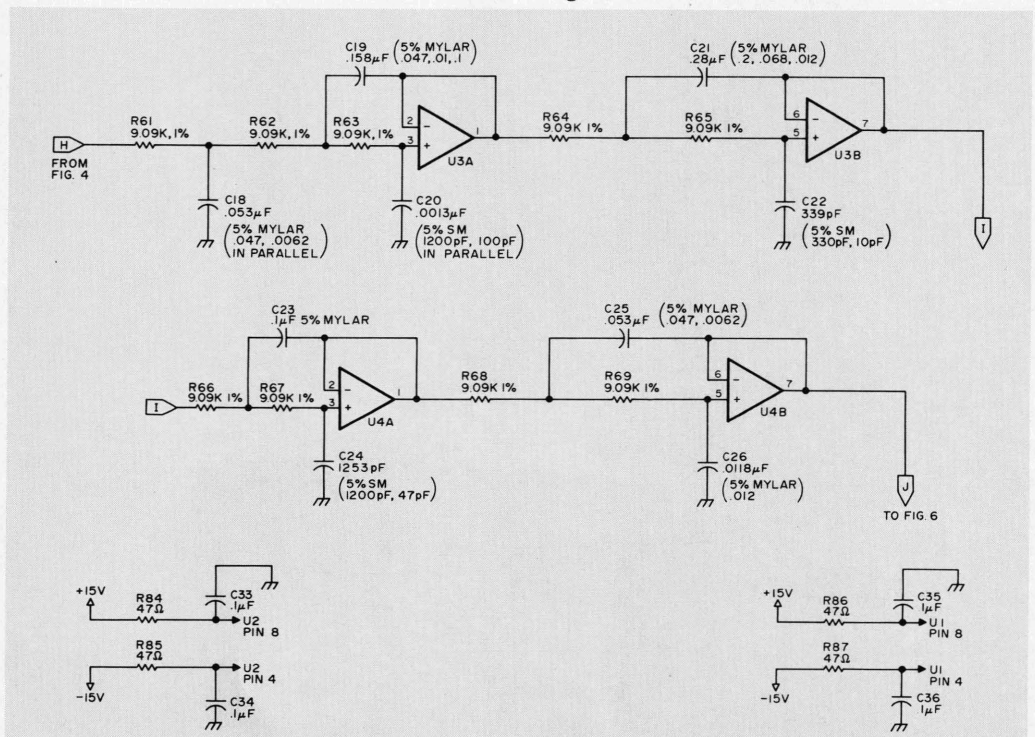


Fig. 5. Video filter. C18-26 are made up from capacitors in parallel.

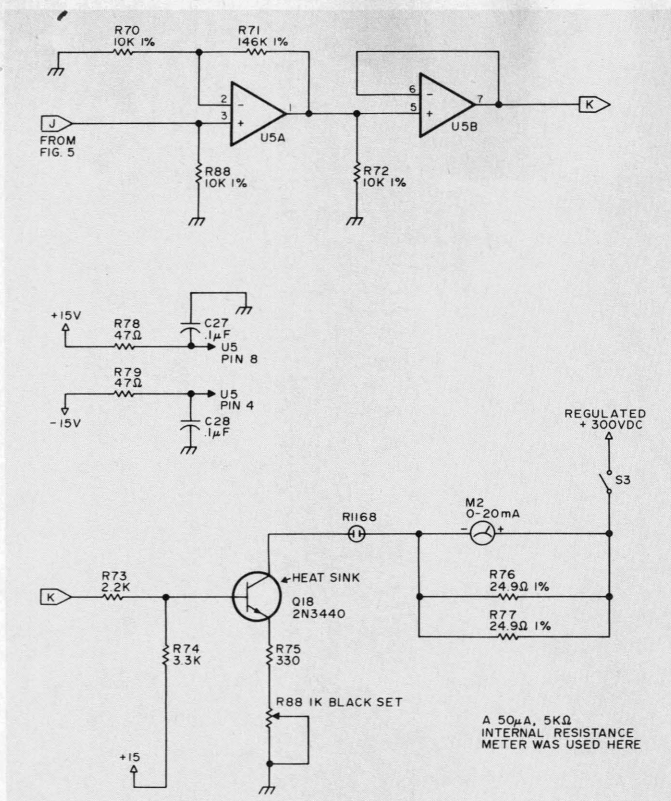
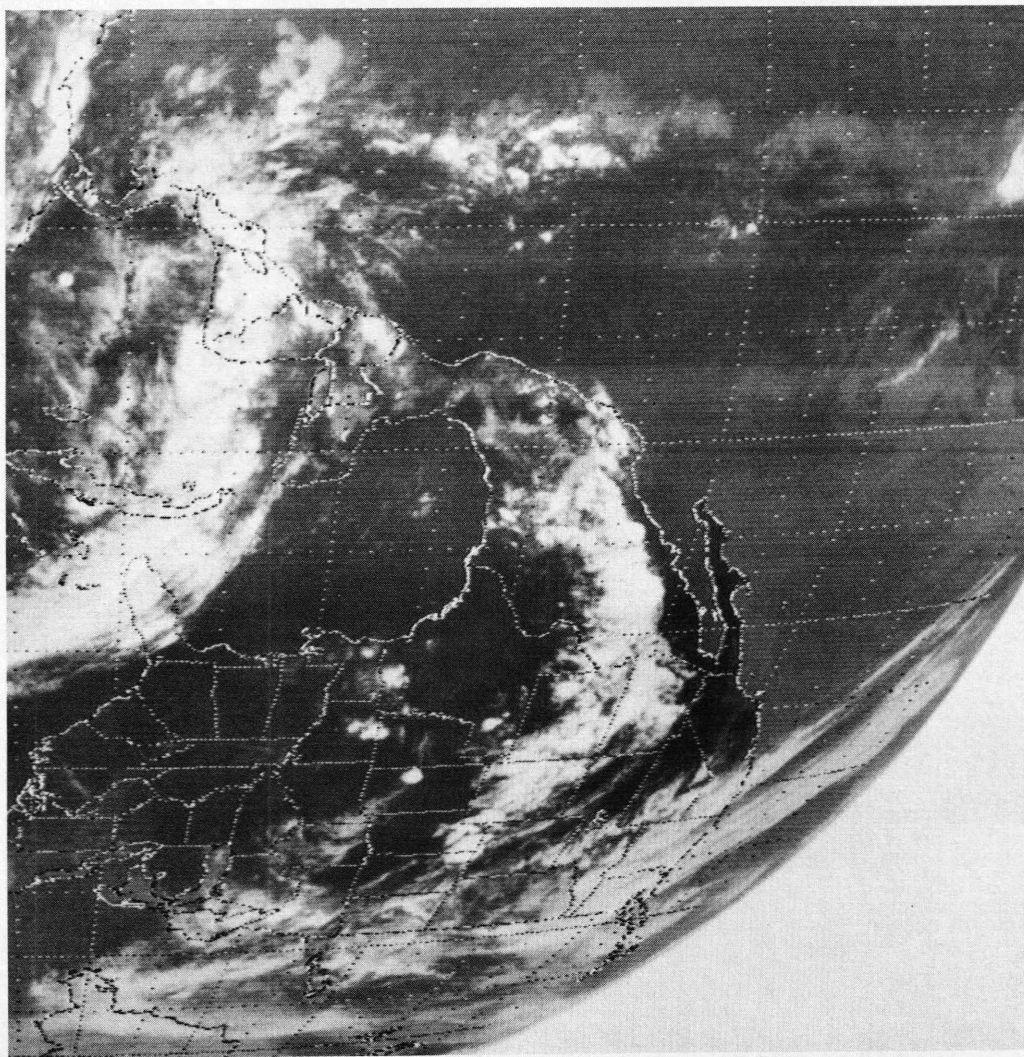


Fig. 6. Lamp driver.



video, point D, is the composite double-sideband AM signal on a 2400 Hz subcarrier. This signal is buffered by U2A and is coupled to the video detector by pot R52. Pot R52 sets the video drive level and is labeled white set. This adjustment controls the white density of the satellite picture.

ICs U1A and U1B comprise the full-wave video demodulator. The demodulated, or rectified, signal at U2, pin 7, contains the subcarrier along with the 0-1600 Hz video information. This rectified video signal is routed to Fig. 5, point H, and becomes the input to the video filter. The video filter is a low-pass, 9th order, .5 dB ripple Chebyshev. It is very flat from 0 to about 1700 Hz and then drops off rapidly

to the noise floor. The 0-to-1700 Hz passband allows the desired video information to pass while presenting a high attenuation to the 2400 Hz subcarrier. Once the subcarrier is removed, you are left only with dc variations corresponding to shades of grey in the satellite picture.

The filter output, point J, finds its way to the lamp driver circuit in Fig. 6. ICs U5A and B accept and amplify the filter output to a level sufficient to drive the 2N3440 lamp driver transistor. The video information is used to control the current through Q18 and, in turn, modulate a suitable light source. In this case, an R1168 glow-tube modulator is used. The polarity of the signal on Q18 is such that black, minimum signal, produces the most lamp current, while a white level decreases lamp current. Pot R88 in the emitter circuit of Q18 is used to set the desired black current.

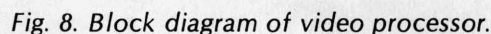
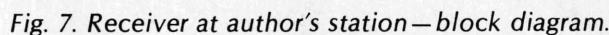
It should be apparent that, with the polarities given, the circuit is set up for positive printing on photographic enlarging paper. I get best results on Kodabrome RC-FH paper. The milliammeter in the collector circuit of Q18 measures lamp current. It is used for calibration of R52 and R88. Finally, a calibration procedure is needed for the video processor. The few simple adjustments are given next.

Calibration

First of all, check the wiring! If you have come this far, you have probably already vowed never again to build another satellite project and are certainly in no mood for an explosion. After all, this thing takes some time to wire. A good point to start calibration is by measuring the p-p output level of your receiver discriminator. You should have at least 250-500

Once you have established the level from your receiver, find a 2400 Hz sine wave signal source and set its output to the same level. Switch S2 to real time, and put the test signal into J1. Set your agc recovery switch to fast. With a scope connected to point D in Fig. 2, adjust R22 slightly and wait. The wait period is to give the agc circuit time to recover. Note the meter reading on M1 as you adjust R22. The meter should not be pegged, but should be somewhere between 10 to 30 μ A. Finally, after a few patient minutes, you should have a 3.3 V p-p signal at point D.

Next, the white and black set pots are adjusted. First, set S2 to real time and disconnect anything attached to J1. Set R52 and R88 fully counterclockwise. Keep your agc recovery switch in the fast mode. Turn on the lamp with S3. Advance pot R88 clockwise until 9.5 mA is reached on M2. This is the black current set. Reconnect the test signal and turn R52 clockwise until M2 falls within 1.5 and 2.0 mA. This is the white current set. During live reception, the lamp current will average out at 3-5 mA or so. With these current settings and about a 40-second development



Parts List

C1	560 pF, 5%, silver mica	R7	270, .25 W, 10%	R62	9.09k, 1%, .25 W
C2	25 uF, 25 V	R8	15k, .25 W, 10%	R63	9.09k, 1%, .25 W
C3	150 pF, 5%, silver mica	R9	3.3k, .25 W, 10%	R64	9.09k, 1%, .25 W
C4	2 uF, 25 V	R10	22k, .25 W, 10%	R65	9.09k, 1%, .25 W
C5	.1 uF, 50 V, monolithic	R11	1.5k, .25 W, 10%	R66	9.09k, 1%, .25 W
C6	2 uF, 25 V	R12	390, .25 W, 10%	R67	9.09k, 1%, .25 W
C7	2 uF, 25 V	R13	1k, .25 W, 10%	R68	9.09k, 1%, .25 W
C8	2 uF, 25 V	R14	1k, .25 W, 10%	R69	9.09k, 1%, .25 W
C9	4.0 uF, 200 V, mylar	R15	1k, .25 W, 10%	R70	10k, 1%, .25 W
C10	.1 uF, 50 V, monolithic	R16	1k, .25 W, 10%	R71	146k, 1%, .25 W
C11	100 uF, 50 V	R17	680, .25 W, 10%	R72	10k, 1%, .25 W
C12	100 uF, 50 V	R18	680, .25 W, 10%	R73	2.2k, .25 W, 10%
C13	10 uF, 50 V	R19	100k, .25 W, 10%	R74	3.3k, .25 W, 10%
C14	3900 pF, 5%, silver mica	R20	5.6k, .25 W, 10%	R75	330, .5 W, 10%
C15	47 pF, 5%, silver mica	R21	1k, .25 W, 10%	R76	24.9, .5 W, 1%
C16	.0047, 100 V, mylar	R22	500 pot	R77	24.9, .5 W, 1%
C17	.01 uF, 50 V, monolithic	R23	200 meg, 5%	R78	47, .25 W, 10%
C18	.053 uF	R24	10k, .25 W, 10%	R79	47, .25 W, 10%
C19	.158 uF	R25	100k, .25 W, 10%	R80	47, .25 W, 10%
C20	.0013 uF	R26	150k, .25 W, 10%	R81	47, .25 W, 10%
C21	.28 uF	R27	1 meg, .25 W, 10%	R82	47, .25 W, 10%
C22	339 pF	R28	10k, .25 W, 10%	R83	47, .25 W, 10%
C23	.1 uF	R29	15k, .25 W, 10%	R84	47, .25 W, 10%
C24	1253 pF	R30	10k, .25 W, 10%	R85	47, .25 W, 10%
C25	.053 uF	R31	270, .25 W, 10%	R86	47, .25 W, 10%
C26	.0118 uF	R32	33, .25 W, 10%	R87	47, .25 W, 10%
C27	.1 uF, 50 V, monolithic	R33	270, .25 W, 10%	R88	1k pot, 2 W
C28	.1 uF, 50 V, monolithic	R34	33, .25 W, 10%	Q1	2N2270
C29	.1 uF, 50 V, monolithic	R35	1.2k, .25 W, 10%	Q2	2N404
C30	.1 uF, 50 V, monolithic	R36	1.2k, .25 W, 10%	Q3	2N404
C31	.1 uF, 50 V, monolithic	R37	1.5k, .25 W, 10%	Q4	2N404
C32	.1 uF, 50 V, monolithic	R38	4.7k, .25 W, 10%	Q5	2N1605
C33	.1 uF, 50 V, monolithic	R39	27, .25 W, 10%	Q6	2N1671A
C34	.1 uF, 50 V, monolithic	R40	220k, .25 W, 10%	Q7	2N1671A
C35	.1 uF, 50 V, monolithic	R41	220k, .25 W, 10%	Q8	2N1131
C36	.1 uF, 50 V, monolithic	R42	8.2k, .25 W, 10%	Q9	2N3819
CR1	1N816	R43	470, .25 W, 10%	Q10	2N2270
CR2	1N816	R44	2.2k, .25 W, 10%	Q11	2N1605
CR3	1N459A	R45	120k, .25 W, 10%	Q12	2N3134
CR4	1N459A	R46	18k, .25 W, 10%	Q13	2N404
CR5	1N459A	R47	470, .25 W, 10%	Q14	2N2270
CR6	1N816	R48	1.5k, .25 W, 10%	Q15	2N2270
CR7	1N816	R49	4.7k, .25 W, 10%	Q16	2N404
CR8	1N816	R50	470, .25 W, 10%	Q17	2N404
CR9	1N816	R51	10k, .25 W, 10%	Q18	2N3440
CR10	1N2484	R52	10k, 10-turn precision pot	U1-5	MC1458 op amp
CR11	1N2484		panel mount	K1-2	12 V dc reed relays
CR12	1N914	R53	10k, 1%, .25 W	S1	DPDT toggle
CR13	1N914	R54	20k, 1%, .25 W	S2	1-pole, 3-position rotary
CR14	1N2484	R55	10k, 1%, .25 W	S3	1-pole, 1-position ceramic rotary
R1	47k, .25 W, 10%	R56	10k, 1%, .25 W	M1	0-50 uA Simpson meter model 1212
R2	50k pot	R57	10k, 1%, .25 W	M2	0-20 mA or 50 uA with shunt, 5k int. resistance
R3	15k, .25 W, 10%	R58	5k, 1%, .25 W		
R4	8.2k, .25 W, 10%	R59	10k, 1%, .25 W		
R5	8.2k, .25 W, 10%	R60	10k, 1%, .25 W		
R6	820, .25 W, 10%	R61	9.09k, 1%, .25 W		

time on the picture, I get excellent results.

Now for the moment you have been waiting for: Switch S2 to the test mode. If you have the test-pattern generator working properly, set its output level pot to give the same agc level on M1 you normally see in real time. This will

protect the agc circuit from accidental saturation. Now, switch on the lamp, start your drum recorder, and make a test photo. If everything is fine, you should see 12 very nice steps from black to white.

A final note here: Since this article has only dealt with the electronics in the

video processing, it is assumed that you already have a facsimile machine. I feel that the *Weather Satellite Handbook* and articles appearing in *73 Magazine* should be explored before attempting construction of a satellite facsimile recorder. I will be more than happy to share

some of my automatic picture phasing circuits and phased locked horizontal sync circuits with anyone interested in automating their drum-type facsimile machine. For any help or requests for additional information, please enclose an SASE to speed the return. ■

an alcohol lamp.

After the lead melted we sprinkled it with powdered sulphur. The sulphur flamed and combined with the lead to form lead sulfide (PbS), a hard crystalline mass. You and your class could experiment with different mixes until you got a "hot" crystal. We used to mount the crystal in Wood's metal that was salvaged from a blown fuse cartridge or plug. Don't forget that the crystal set works best when its coil is wound around a genuine Quaker Oats oatmeal box.

SIMPLE RF FIELD MONITOR

I enjoy flying radio-controlled

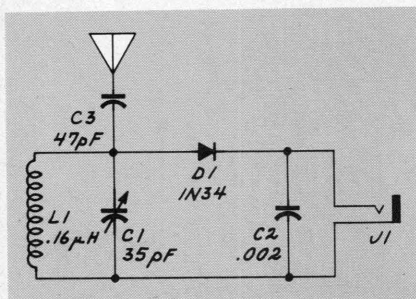


FIG. 3

model airplanes. I use AM digital/proportional control on Channel 44 (72.670 MHz). I need to know if someone is using that channel before I put my plane in the air. Can you give me the circuit of a simple monitor for that frequency?—B. G., Evergreen Park, IL.

A simple crystal receiver should do the trick. The circuit in Fig. 3 shows a suitable circuit. Coil L1 should be wound from eight turns of 18-gauge enameled wire spaced two inches apart on a 1/2-inch form. Use high-impedance phones (2000 ohms or more) for best sensitivity.

FM NOTCH FILTER

I want to notch-out a local FM station at 103.9 MHz so that I can pick up a distant station at 103.7 MHz. I feel that the solution lies in using op-amp active filters. Can you help me?—N. F., Clifton Heights, PA.

I don't feel that using op-amp active filters is the most practical way to go.

As an alternative, an inexpensive and highly effective notch filter can be built merely from a section

of coaxial transmission line that is connected across the lead-in at, or

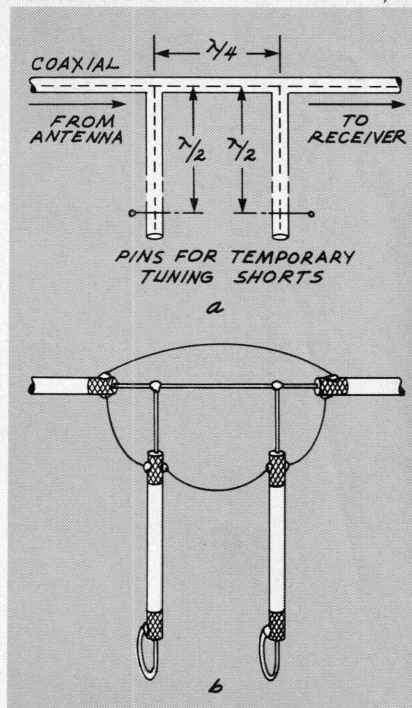


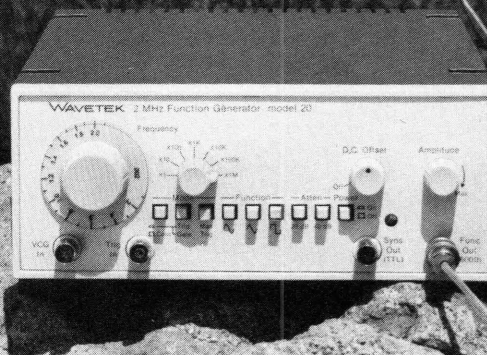
FIG. 4

close to, the receiver's antenna input. With careful measurement continued on page 99

WAVETEK®

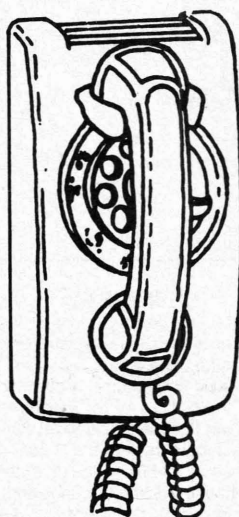
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R-E Engineering Admart

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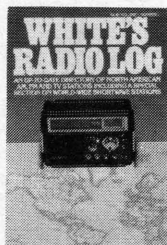
Size	Secondary Volts/Amps	Price each*		
		1	2-4	5-24
25VA	2x8.5V/1.47A	\$24.30	\$20.40	\$17.90
OD = 2.4"	2x12V/1.08A	24.30	20.40	17.90
H = 1.2"	2x15V/0.87A	24.30	20.40	17.90
Wt = 0.7 lb	2x18V/0.71A	24.30	20.40	17.90
	2x24V/0.54A	24.30	20.40	17.90
180VA	2x12V/7.50A	\$36.60	\$31.20	\$27.80
OD = 3.9"	2x15V/5.90A	36.60	31.20	27.80
H = 1.9"	2x18V/5.00A	36.60	31.20	27.80
Wt = 3.0 lb	2x24V/3.58A	36.60	31.20	27.80
1000VA	2x50V/10.00A	\$91.70	\$78.40	\$69.00
OD = 6.3"	2x65V/7.69A	91.70	78.40	69.00
H = 2.4"				
Wt = 12.1 lb				

Dual secondaries may be used in parallel or in series.
*FOB Factory. (Includes mounting washer and rubber pads).
Terms: VISA, MasterCard, COD or money orders—UPS charges added.

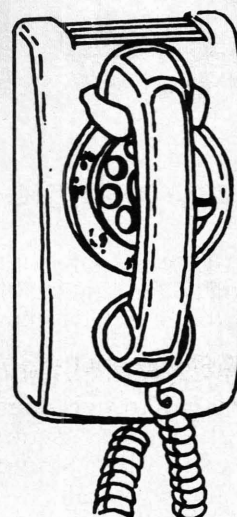
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ASK R-E

continued from page 14

and assembly, it is possible to build a simple filter that provides as much as 30 dB of attenuation at the desired frequency. If you wish, two filter sections can be used to sharpen the notch and provide up to 80 dB of attenuation.

The operation of a transmission-line stub filter is based on some unique characteristics of a transmission line. A section of line that is one-half wavelength long and shorted at one end will appear as a short at the frequency of interest at the other end of the line. The basic idea is shown in Fig. 4-a. Purists

may insist that the spacing between stubs be $\frac{1}{4}$ wavelength, but we have found through experimentation that, when using those filters to treat TVI problems, the stubs can be used as close together as is mechanically practical.

The physical length of a half-wave stub will be shorter than its electrical length because of the velocity factor, or propagation constant, of the cable from which the stub is built. The velocity factor, K, for RG-59/U 72-ohm coax is 0.66, and it's 0.83 for 300-ohm twin lead. The formula for a half-wave stub of coaxial line is:

$$L = (5906 \times K)/f$$

In that equation, L is the length in inches, K is the velocity factor of

the transmission line, and f is specified in MHz. So, to build a 103.9-MHz notch filter, $L = (5906 \times 0.66)/103.9 = 37.5$ inches.

Build the filter as shown in Fig. 4-b, with the stubs an inch or so longer than the calculated value. Connect the filter in the line between the antenna and the receiver, and then tune in the unwanted station. Now, starting at the far end of the stub, use a needle or a sharp pin to short the outer braid to the inner conductor in $\frac{1}{4}$ -inch steps until you find the point of maximum attenuation. Mark that point, increase the stub length about $\frac{3}{4}$ inch and cut off the excess. Strip off $\frac{1}{2}$ inch of insulation and solder the core to the braid. That should do it! **R-E**

Ask R-E

PC-BOARD GROUND PLANE

I'm familiar with the term *ground plane* as applied to antennas, but not to PC boards. What is the significance and importance of ground planes on PC boards? Can you suggest additional reading on the subject?—A. M. W., Nathrop, CO.

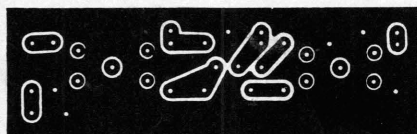


FIG. 1

A ground plane, as shown in Fig. 1, is a continuous, un-etched area of copper that is connected to circuit ground. Ground planes are used to minimize stray inductive and capacitive coupling and to eliminate ground loops in low-level, low-noise audio circuits, in HF and VHF circuits, and in digital circuits.

The ground plane may be on the wiring side of a PC board, the component side, or both. When it is on the component side of the board, the ground plane is generally unbroken, except for clearance around the component-terminal holes. When the ground plane is on the plated wiring side of the board, sensitive circuit lines are run close by, and parallel to the ground plane; alternatively, signal lines may be routed through the ground plane.

Our illustration shows the PC board pattern for a filter used in a 2-meter transmitter. The ground plane is on the wiring side of the board.

For more information on using and designing ground planes, refer to *The Design & Drafting of Printed Circuits* by Darryl Lindsey, published by Bishop Graphics,

Inc., West Lake Village, CA. Also see *Generation of Precision Artwork for Printed Circuit Boards* by Preben Lund, published by John Wiley and Sons, New York, NY.

SPEAKERPHONE INFORMATION

I saw the article, "New All-In-One Speakerphone," in the November, 1978 issue of Radio-Electronics, but I've been unable to contact either the author, Mr. Camenzind, or Tridar Corporation, the maker of the phone. I'd like to get a complete diagram and parts list, if they're available. The Tridar phone appears to include circuits that overcome the switching problem that develops when both parties on the phone line speak at the same time.—M. L. A., Philadelphia, PA.

Telephone technology has advanced quite a bit in the eight or so years since the Tridar speakerphone was developed. I'm sure that you can find a satisfactory speakerphone in a well-stocked telephone-supply store.

By the way, Motorola recently announced a new IC: the MC34018 Voiceswitch Speakerphone. That IC incorporates all necessary amplifiers, attenuators, and control functions to produce a hands-free speakerphone system. For a data sheet and circuit application diagrams, contact Mr. Ron Hlavinka, Motorola Semiconductor Products, P.O. Box 20912, Phoenix, AZ.



MEASURING SPEAKER IMPEDANCE

I am in auto-sound servicing and installation, and I have accumulated quite a few speakers. I don't know the impedance of many of those speakers. I know that I can't measure impedance with just a multimeter, but I don't know which instruments

WRITE TO:

ASK R-E

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to use. I have most of the usual audio service instruments. Can you help me?—R.H., Hartsville, SC.

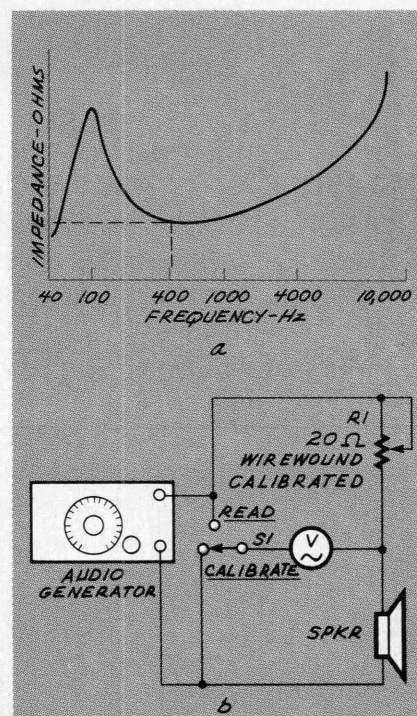


FIG. 2

The impedance of a loud-speaker varies with frequency along a curve like that shown in Fig. 2-a. The speaker's impedance can be measured with an audio generator, a variable resistor, and a high-impedance AC voltmeter. Wire the equipment as shown in Fig. 2-b, and set the generator's output for 400 Hz. With S1 in the CALIBRATE position, adjust the generator's output attenuator to give a full-scale reading on the meter. Then throw the switch to READ and adjust R1 until the meter reads the same in that position. The speaker's impedance then equals R1.

continued on page 12

BUILD THIS

THUNDERSTORM ALARM



Don't be caught unawares by the sudden arrival of a thunderstorm with its accompanying wind and rain. This simple radio accessory gives an early warning.

CALVIN R. GRAF

THUNDERSTORMS, AND THEIR ACCOMPANYING strong winds, rain, and possible hail, can make their appearance rather suddenly sometimes. This is especially so in spring and summer months, but they can actually sneak up on you at almost any time in some parts of the country. When camping out, fishing, picnicking, or just relaxing at home, it is important to know of any severe weather that might be approaching the local area. This is of special interest to those who have to conduct outdoor operations such as construction workers, farmers, and ranchers. Campers, away from their vehicles, can be warned to seek higher ground in case of flash floods.

The thunderstorm activity indicator described in this article will alert you to an approaching electrical storm through the flashing of two light emitting diodes (LED's) and the sounding of an audio alarm. The activity indicator is connected to the earphone audio output jack of a pocket transistor radio or connected across the speaker terminals of any radio receiver. The radio is then tuned to a clear spot near the upper end of the broadcast band (1600 kHz) where there are no stations being received. An AC power supply with 9-volt DC output can be used to operate the radio at home. With this supply, the receiver can be left on continuously and the receiver will consume little power but will provide an alert no matter the time of day or night a storm may appear. The AC-operated supply is inexpensive and can be purchased at any local radio store. A volume control is provided so that the audio alert level may be adjusted or turned down completely. A visual alert is still provided, however, by the continuous flashing of the LED's as a storm appears.

The circuit diagram shows how the alarm is connected to the receiver. Transformer T1 is a small transistor radio

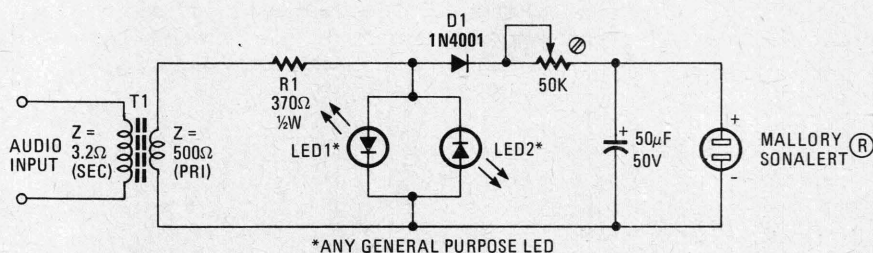


FIG. 1—THUNDERSTORM ACTIVITY indicator and alarm. The audio input terminals are connected to the speaker terminals of any AM broadcast radio receiver.

output transformer connected in reverse. It is used to raise the audio voltage across the loudspeaker (3.2 ohms) to a level that will cause the LED's to operate properly (500 ohms). Resistor R1 serves as a current limiting resistor for the LED's so that the voltage drop across them never exceeds a nominal 1.6 to 1.7 volts. The LED's are connected in reverse polarity parallel so that one will conduct in the forward (positive) direction of the audio signal and other LED will conduct in the reverse (negative) direction of the audio.

Diode D1 is used to rectify the alternating audio voltage so that only pulsating DC is applied to the Sonalert as its polarity markings must be observed. The Sonalert emits a pleasant 2900 Hz signal when the applied voltage is a nominal 1 volt DC. The capacitor charges up on the sharp noise impulses that occur each time there is a lightning flash. When the voltage across the capacitor rises to a value close to one volt, the Sonalert will emit a long "ping". The capacitor thus serves as an integrator and stores up lightning flashes before it causes the Sonalert to sound forth. In this manner, short noise transients on the power line that are radiated from light switches, air conditioners and the like, do not cause the Sonalert to sound. Output from the alarm is also dependent on the setting of the receiver volume control and it will sound out with a normal room level setting.

When a thunderstorm is 10 to 20 miles away, the audio output from the radio due

to atmospheric disturbances will cause the LED's to flash and the Sonalert to sound. As the thunderstorm approaches the local area, thunder may be heard following the "ping" of the Sonalert. Knowing that sound travels one fifth of a mile per second in air, the exact distance to the storm area can be calculated by counting seconds from the time the ping is heard until the thunder is heard. If you count to five, the storm is one mile away, and so forth. When the Sonalert sounds continuously, the electrical storm and accompanying rain are very nearby.

The approximate direction to the storm can be determined by "aiming" the receiver's antenna toward the storm area that produces maximum audio output from the Sonalert. The storm passage through the local area can be followed by plotting the relative bearing against time. Keep the volume level constant.

Remember, as the storm approaches, light intensity of the LED's and the sound duration from the Sonalert will increase. As the storm recedes, the relative levels of both light and sound will drop. The storm passage may last from 30 minutes to several hours. With a little experience, you will soon learn to recognize whether it is going to rain or not, in spite of what the weather man may say! (If you live in the cyclone or tornado belt consider using the Stormwarn alarm along with a tornado alert device based on light flashes on a blank TV raster.—Editor)

R-E

*Couple a function generator to an oscilloscope
to form a versatile instrumentation set for
testing systems and components.*

ROBERT KRAL

FUN WITH FUNCTION GENERATORS

THE FUNCTION GENERATOR IS A versatile test instrument that usually functions as a signal substitution source. This article explains how the function generator, when coupled to an oscilloscope, becomes an even more versatile instrumentation set for testing and evaluating systems and components.

The function generator is an instrument that permits the introduction of its signals at specific points in a circuit. This characteristic makes it useful as an audio generator for testing, troubleshooting, and evaluating audio products as well as conventional electronic circuit components.

The oscilloscope paints the "signatures" of electronic components on its screen and a digital multimeter (DMM) makes the routine electrical measurements necessary to perform the tests discussed in this article. In addition to its role in testing, this instrument suit is very effective teaching tool for ex-

plaining the behavior of both active and passive components to students.

What do you need?

To perform the tests and demonstrations described in this article, the function generator, oscilloscope and DMM should have the following minimum specifications:

- Function generator—2-MHz internal sweep
- Oscilloscope—two-channel, 20-MHz with X-Y operation
- Digital multimeter—capable of making the basic electrical measurements with a basic DC accuracy of at least 1.0 %

Manufacturers of function generators position the controls and connecting plugs in different ways on the instruments that are designed to fill various market segments. Therefore, there really are no standard signal generators. If your function generator differs from the one shown in Fig. 1, you might want to consult your manual to re-

view the functions of all of its controls.

Fortunately, most signal generators have the same basic controls and input terminals that can be interconnected in the same way. The leading signal generator specifications are: frequency range, sinewave distortion, types of waveforms generated, sweep range, and signal output level.

Selecting a function generator

If you do not own or have access to a signal generator, there are some points to consider when selecting one. Most stock function generators have a frequency bandwidth that extends from a fraction of a hertz to 10 MHz or higher. Most can produce sine, square, and triangle waves. On many instruments, variable symmetry control, also called duty cycle adjustment, allows squarewave pulses to be narrowed to simulate clock pulses.

Amplitude or frequency mod-

ulation and internal linear or logarithmic sweep are available on some instruments. Separate TTL or CMOS pulse outputs might be available, providing exact preset voltage levels required for a specific logic family. Some function generators contain a frequency counter that will give a direct readout of output frequency.

The tests described in this article can be performed on a basic economy model function generator suitable for most electronic hobbyists and most routine electronics maintenance and circuit design. None of the tests described here require a bandwidth in excess of 2 MHz because they are limited to the audio range.

The sinewave distortion specification even for a basic function generator should be less than 1%, and its output should be variable, up to 20 volts peak-to-peak into an open circuit. A built-in counter is desirable feature but is not required for setting the output frequency. An internal linear or log sweep will be needed for making audio measurements. Instruments that meet these requirements typically sell in the \$300 to \$400 range.

Frequency-response

A function generator is not a constant voltage source. It typically has an output impedance of 50 ohms. Consequently, if you want to test a low-impedance component such as a speaker connected to the generator, the output of the function generator will vary with the impedance of the load.

To correct this condition, either add a series resistor or set up the function generator to drive an audio amplifier which, in turn, will feed the speaker. The low output impedance of the audio amplifier makes the function generator a constant voltage source.

Set up your instruments as shown in Fig 2. Select X-Y operation on your oscilloscope. Temporarily connect the function generator's output directly to the oscilloscope's vertical Y-axis input (typically channel 1). Con-

nect the generator control voltage (GCV) output to the oscilloscope's horizontal X-axis input (typically channel 2).

Select DC coupling on both input channels. Display only Channel 1. While the sweep is disabled, adjust the function generator to the desired maximum (ending) sweep frequency. A vertical line should appear on the screen. Adjust the oscilloscope's channel 1 position and voltage controls so that the line is centered vertically and is positioned to the far right, horizontally.

Adjust the amplitude controls so that the trace nearly fills the screen vertically. Then set the function generator's frequency lower. The vertical line should move to the left. Tape a piece of paper to the CRT screen and mark the position of the vertical trace at the various frequencies that you will use in your tests.

Tune the function generator back to the end (maximum) sweep frequency. Engage the sweep width adjustment, and set it for the maximum. Adjust the sweep width and frequency controls to obtain the desired

start and stop frequencies. The generator will now sweep between these two frequencies.

Now connect the component to be tested. It could be an amplifier, passive or active filter, loudspeaker, or microphone. Adjust the controls as needed so that the oscilloscope display looks like that shown in Fig. 3.

Impedance measurements

Figure 4 shows the setup for making impedance measurements on speakers and passive filters. Connect the GCV output of the function generator to the oscilloscope's X-input. With the oscilloscope in the X-Y mode, the oscilloscope's Y-input is the voltage across the speaker.

With the function generator out of the sweep mode, connect a 2-kilohm resistor in series with the function generator's output. Position the waveform near the center of the oscilloscope screen, and adjust its amplitude.

Adjust the output frequency of the generator until the vertical line reaches its maximum length. This is the speaker's resonant frequency. You can sub-

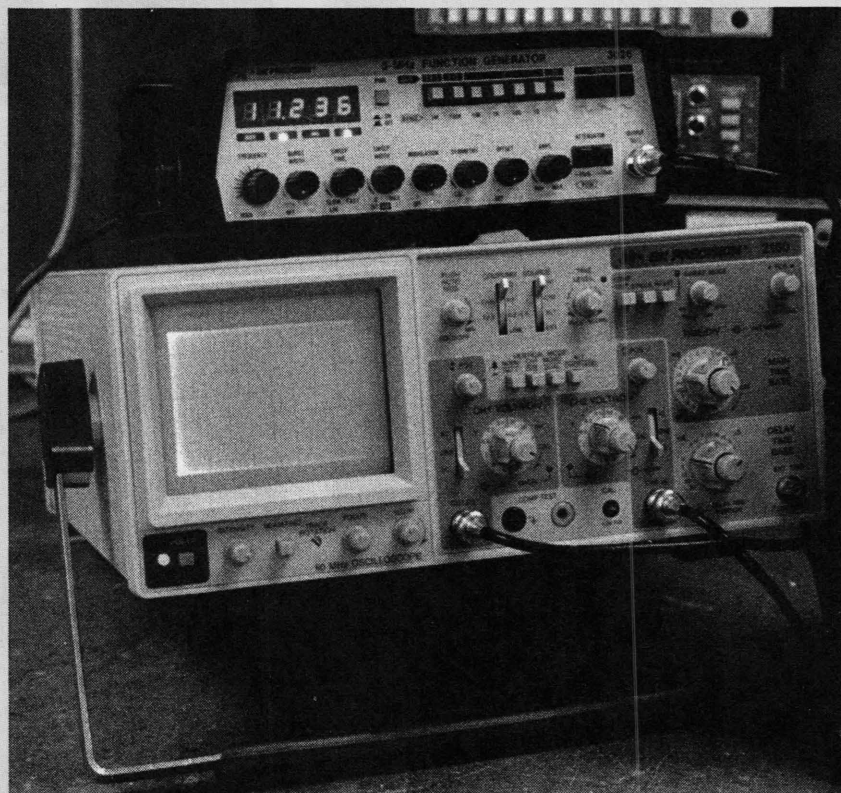


FIG. 1—A FUNCTION GENERATOR/OSCILLOSCOPE SETUP suitable for making the tests described in this article.

stitute a potentiometer for the speaker to determine the value of the speaker's impedance at any frequency. Adjust the potentiometer to produce a line with the same vertical length that the speaker produced. Then measure the resistance value of the potentiometer with the DMM set on the ohms scale.

An enclosed audio woofer (low-frequency driver) in a non-vented (sealed) enclosure has one impedance peak at system resonance as shown in Fig. 5. The height of the peak indicates damping. The higher the peak, the lower the damping or control at resonance. By contrast, ported (bass reflex) speaker systems have two impedance peaks, one on either side of system resonance. Two resonance peaks in a sealed box system indicate an audio leak or loose internal baffle. Resonance frequencies of midrange and high-frequency drivers are indicators of the low, frequency limits of those transducers.

Amplifier transient response

Because of the high harmonic content of square waves, tests made with them supplement sinewave frequency response measurements for amplifiers with wide bandwidths. Set up the instruments as shown in Fig. 6. If the output impedance of the function generator does not match the input impedance of the device under test, ringing will occur. If the device you are measuring has a high-impedance input, parallel the input with a 50-ohm resistor. That value is equivalent to the generator's output impedance.

Set the function generator for a triangular wave output, and adjust that output so no clipping occurs. Then switch to squarewave output, adjust the oscilloscope, and compare the squarewave output from the amplifier with the square wave produced by the generator.

Figure 7 shows normalized drawings of typical waveforms. In waveform A, for example, the frequency distortion (or amplitude reduction of the low-frequency component) occurs without a phase shift. In wave-

form I, the low-frequency phase shift shows the effects of hum voltage.

Component signature analysis

A signature analyzer has long been the ideal instrument for testing all the components on circuit boards with no power

applied. The analyzer applies an AC signal to the device under test. The analyzer's display, with voltage on the X axis and current on the Y axis, shows the characteristic impedance of the device, or "E-I signature." Short circuits appear as vertical lines and open circuits appear

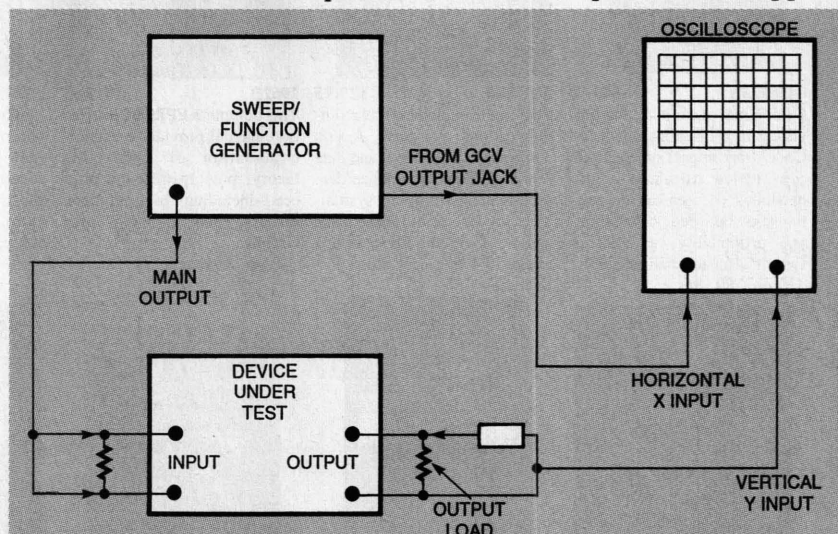


FIG. 2—FREQUENCY RESPONSE TEST SETUP and measurement. Mark the frequencies of interest on the graticule of the oscilloscope with a china marking pencil.

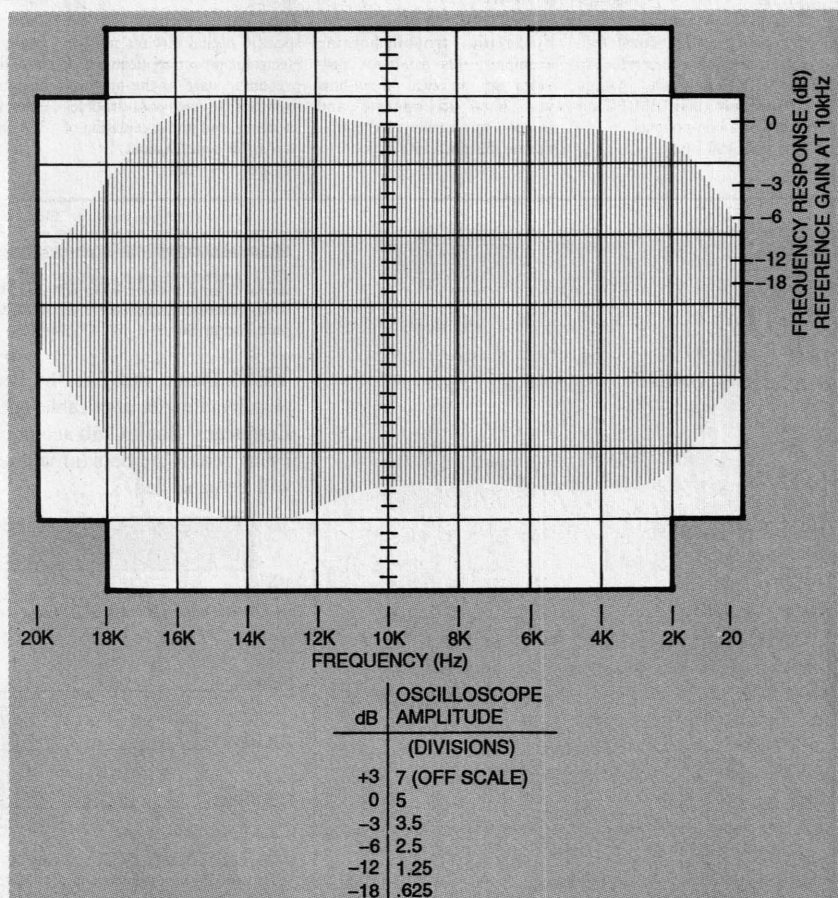


FIG. 3—WAVEFORM FOR A TYPICAL AUDIO AMPLIFIER response display (a) and table of oscilloscope amplitude in graticule divisions vs. decibels (b).

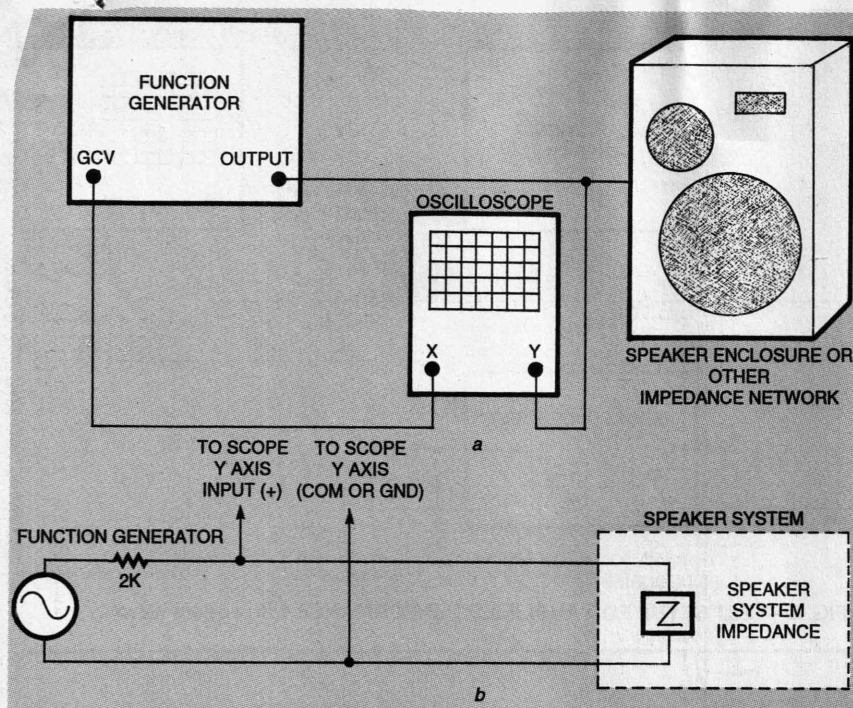


FIG. 4—SETUP FOR TESTING SPEAKER SYSTEMS and impedance networks (a) and equivalent circuit (b).

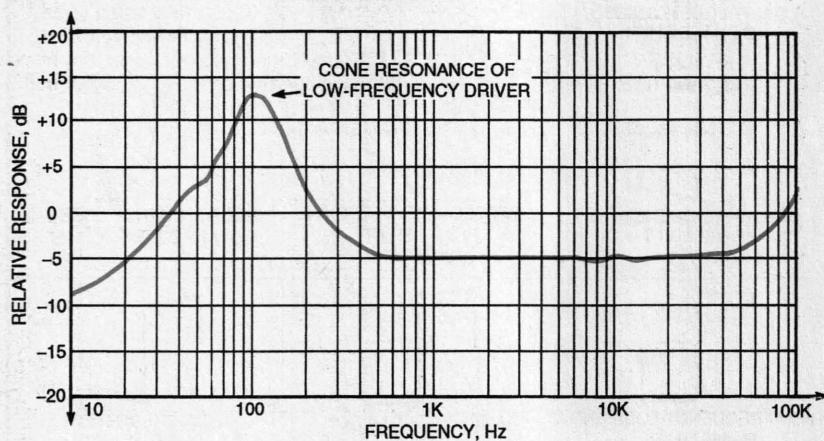


FIG. 5—GRAPH OF RELATIVE RESPONSE vs. frequency for an enclosed, low-frequency speaker showing a single resonance peak.

as horizontal lines. Marginal components can be identified by comparing their waveforms with idealized waveforms of the correct signatures.

Fortunately, you can perform many of the tests best performed on a signature analyzer with a function generator and an oscilloscope and save the \$1000 to \$3000 you would have to spend to buy an analyzer.

With the instrumentation setup described here, the range of applied test voltage will be somewhat limited. Moreover, repeatability will not be optimum because you must readjust frequency and output voltage each

time you perform a test. However, you will become familiar with the E-I test technique without the need to purchase additional instruments.

This setup lacks some of the benefits of the full-fledged signature analyzer such as A-B input switching for rapid good-bad comparisons, multichannel clips for testing the pins of integrated circuits automatically in sequence, and the ability to store component waveforms on a personal computer for comparison purposes. And, of course, without this connection to the PC, you won't be able to generate reports with the

data or store test records.

The test setup for the alternative signature analyzer is shown in Fig. 8. Connect the function generator output across both the component under test and a 51-kilohm series resistor. Fix the oscilloscope's Y input across the 51-kilohm resistor in series with a 2-kilohm resistor, and connect its X input across the component under test. Tie the grounds from the two probes to a common point between the resistor and component under test.

Select the sinewave output from the function generator, and set the voltage and frequency controls for a clear display. Again, set the oscilloscope in the X-Y mode.

Familiarity with the waveforms of good components is essential for the most effective troubleshooting. If you are doing routine production testing, have a sample tested and known fault-free board on hand for comparison with any board that contains faulty components. Note, however, that the signature of a component in a circuit will differ from that of the same component out of a circuit.

Select a test voltage and frequency very carefully to produce the optimal signature for the known good component. The signatures expected will, of course, differ with the type of component being tested. Keep a written record of the voltage and frequency settings for the values you test.

Typical signatures

The signature of a resistor is a sloping line as shown in Fig. 9-a. The slope changes with the value of the resistor. Select a voltage so that the slope of the line for a known good resistor is close to 45° .

The signatures of capacitors and inductors are ellipses as shown in Fig. 9-b. They should nearly fill the oscilloscope graticule. High values of each flatten the ellipse along its horizontal (X) axis, and low values flatten it along its vertical (Y) axis.

Here are some points to keep in mind about inductor and ca-

pacitor E-I signature testing:

- Inductors with the same value can have different signatures because of differences in their core materials or the position of the core with respect to the winding.

- Capacitor leakage that cannot be found with digital multimeter capacitance measuring circuits can be found with signature analysis.

Semiconductor diodes can also be tested with this method. Diode signature analysis will detect defective diodes that are open or nearly open-circuited, short-circuited, or those with high internal leakage. However, the same information can usually be obtained with the diode test function of the DMM.

- Silicon diodes exhibit the easily identifiable backward "L"-shaped signatures shown in Fig. 9-c. Light-emitting diodes have similar signatures except that they have a greater slope in their vertical portions. This slope varies with the emission color of the LED, as shown in Fig. 9-d.

Zener diodes conduct in both directions as shown in Fig. 9-e. If the diode is good, it will have a sharp, well defined knee, provided the test voltage is set high enough to produce it.

Transistors and ICs

When the signature technique is applied to transistors and integrated circuits (analog or digital), the waveforms have diode-like signatures:

- Bipolar junction transistors must be tested across their base-collector, collector-emitter, and base-emitter junctions. Their signatures are similar to a silicon diode in series with a Zener diode. The separating point between the two waveforms represents the base.

- Field-effect transistor (FET) must be tested across their gate-source, drain-source, and drain-gate junctions.

- Silicon controlled rectifiers (SCR) are connected across the anode and cathode, and pulse generator output is connected to the gate. Increase the pulse generator output until the SCR turns on.

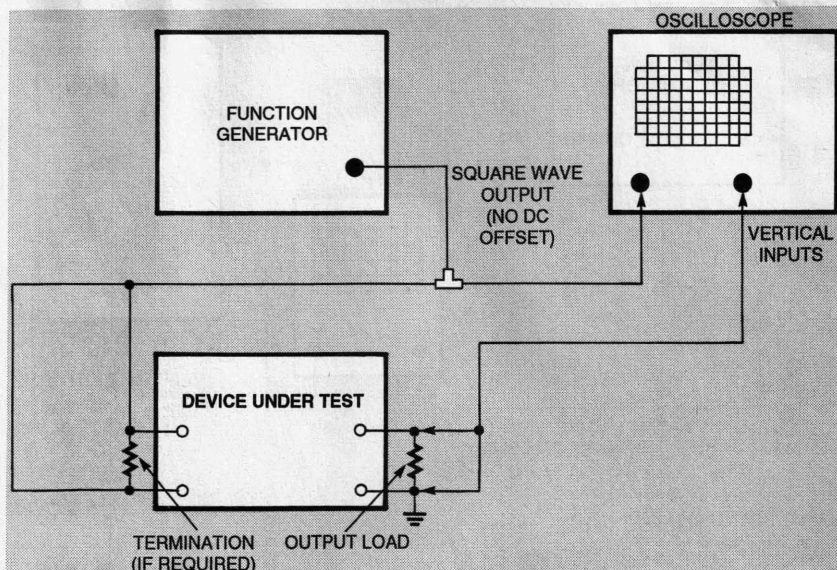


FIG. 6—TEST SETUP FOR AMPLIFIER PERFORMANCE with square waves.

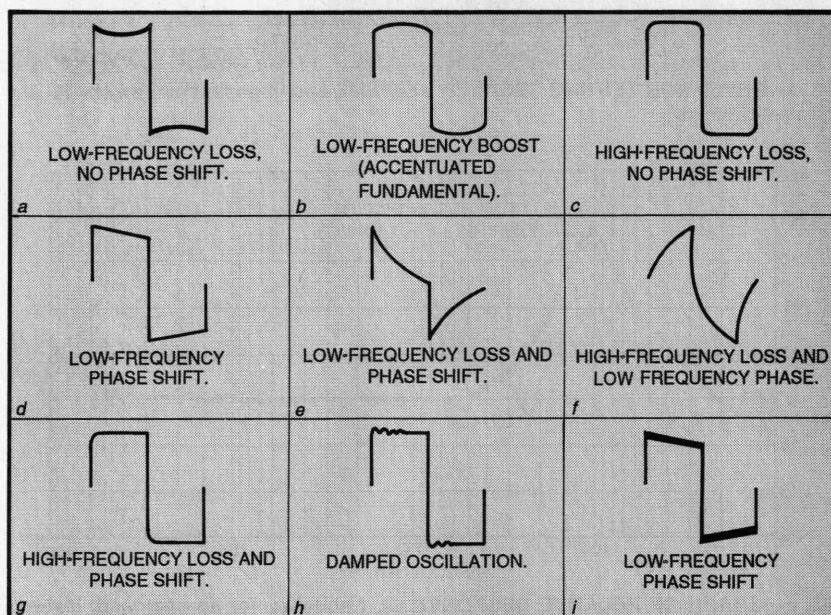


FIG. 7—WAVEFORMS OBTAINED for various test conditions with square waves from the signal generator.

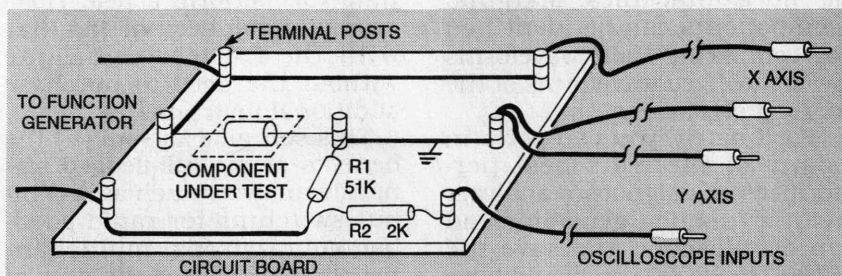


FIG. 8—COMPONENT TEST BOARD that can be hard-wired or put in a project case to simplify the test procedure.

Semiconductor tutorial

Signature analysis is useful in explaining the operating characteristics of semiconductors. For example, in the diode

signature shown in Fig. 9-c, the horizontal distance between the vertical axis on the screen and the vertical part of the sig-

Continued on page 73

RAY MARSTON

THIS ARTICLE PICKS UP WHERE LAST month's article on phase-locked loops left off. Last month the basic principles of the phase-locked-loop were explained and the functional sections of a popular CMOS PLL IC, the CD4046B were described. The circuits presented in that article showed various ways to configure the voltage-controlled oscillator (VCO) section of the CD4046B for practical frequency-dependent applications.

Sound-effects generators

The versatility of the VCO section of the CD4046B IC was explained last month. Its operating frequency can be swept over a very wide range under the control of voltages applied to pin 9. Moreover, its output can be gated on or off with a voltage applied to pin 5. These characteristics make the CD4046B suitable as the principle component in many different circuits for sound-effects generation. Figures 1 to 6 are schematics for sound-effects generators.

The circuit in Fig. 1 can emit a conventional siren sound. It produces a tone that rises slowly to a maximum value when S1 is closed, and falls slowly from that maximum to silence when S1 is opened.

This response is caused by the voltage on C1 that is applied to voltage-control pin 9. It rises exponentially through R1 when S1 is closed, and falls exponentially through R2 when S1 is opened. Resistor R3 ensures that the operating frequency falls to zero when the voltage at pin 9 is zero. The VCO output is AC coupled to the speaker

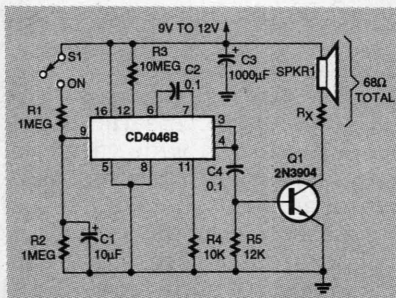
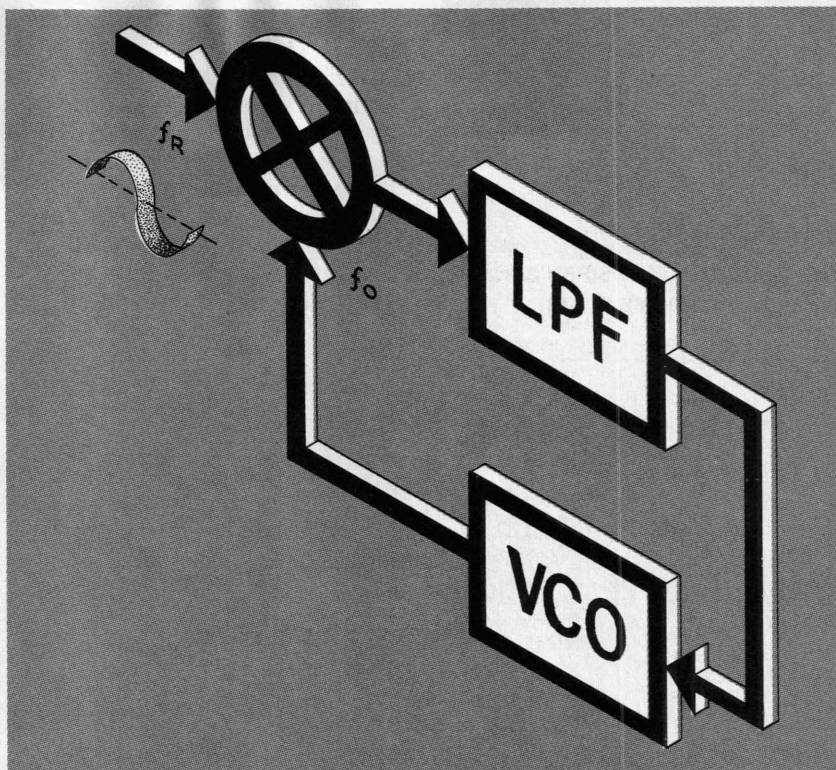


FIG. 1—ELECTRONIC SIREN CIRCUIT based on a CMOS CD4046B PLL IC.



PHASE-LOCKED LOOPS

Experiment with these phase-locked loop circuits, and put them to work in your designs.

through C4 and transistor Q1.

Figure 2 shows how the Fig. 1 circuit can be modified to give a quick response in which the frequency rapidly switches to its maximum value when S1 is closed (as C1 discharges exponentially through R3). Figure 3 shows another circuit modification that generates a "phasor" sound (similar to that heard aboard the starship Enterprise in the Star Trek TV series) when pushbutton switch S1 is closed. In this circuit, the CD4011B IC is configured as an astable multivibrator that is gated through S1. It produces a chain of 4 millisecond pulses at intervals of 70 milliseconds.

Each pulse rapidly charges capacitor C2 through R3 and D2 to produce a high tone which decays slowly as C2 dis-

charges through R5. The cycle is repeated again on the arrival of the next pulse.

Figure 4 shows a different kind of sound-generator circuit that can generate either a pulsed or warble tone, depending on the setting of S2. Pushbutton switch S1 simultaneously enables pin 5 of the CD4046B to gate on the CD4001B astable multivibrator which, in turn, applies a rectangular (alternately fully-high and fully-low) waveform to the IC's pin 9.

In the *pulsed* mode, the VCO generates zero frequency when pin 9 is low, but in the *warble* mode it generates a tone that is 20% lower than the high tone generated when pin 9 is low.

Figure 5 is a circuit that produces a special-effects run-



When PB1 is pressed, C1

The circuit in Fig. 7 is a simple *frequency shift keyed* (FSK) square-wave generator. With the component values shown in



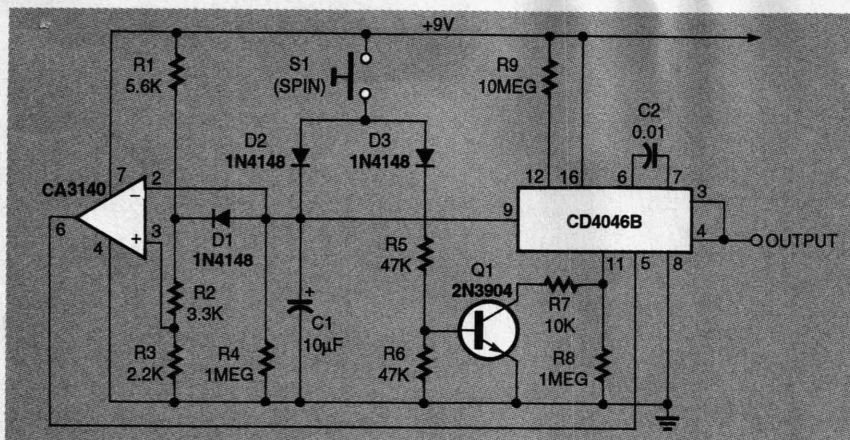


FIG. 6—MODIFIED RUN-DOWN generator circuit.

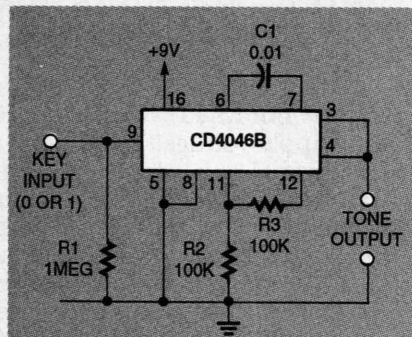


FIG. 7—FREQUENCY SHIFT KEYED (FSK) generator. Logic 0 = 1.2 kHz and logic 1 = 2.4 kHz.

this figure, the circuit will generate a tone frequency of 2.4 kHz when a logic-one signal is applied to pin 9. It will also generate a 1.2-kHz tone when a logic-zero signal is applied to the same point.

The high-frequency tone is set by the values of C1 and R2, and the low-frequency tone is set by the values of C1, R2, and R3. Other frequencies can be obtained by altering those component values.

The circuit in Fig. 8 is a 220-kHz frequency modulation (FM) waveform generator. The internal Zener diode at pin 15 of the CD4046B provides a stable power supply for the CD3140 op-amp, which is configured as a multiply-by-20 inverting AC amplifier. It has a quiescent bias of about 2.6 volts applied to its non-inverting A input at pin 3 through R2 and R3.

As a result, output pin 6 of the op-amp produces an approximate 2.6-volt potential that is amplitude modulated with an amplified ($\times 20$) version of the audio frequency input signal. This output is applied to volt-

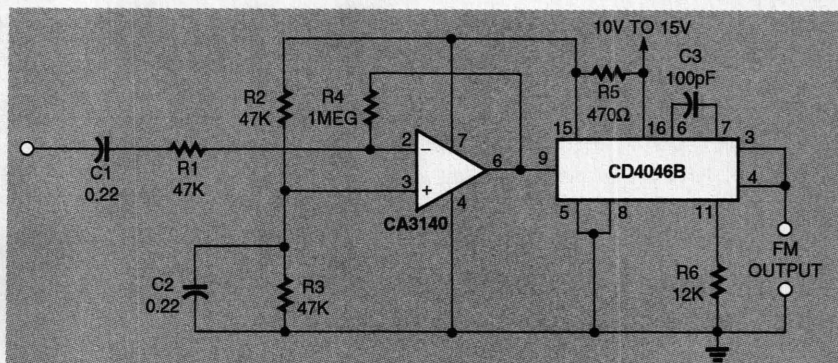


FIG. 8—FREQUENCY MODULATOR CIRCUIT generates 220 kHz.

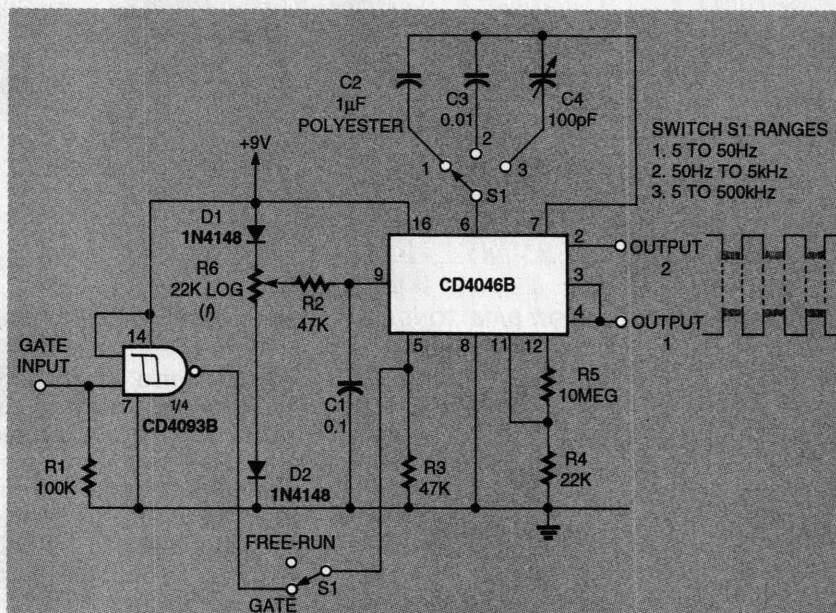


FIG. 9—UNIVERSAL CLOCK/SQUAREWAVE generator.

age-control input terminal pin 9 of the CD4046B's VCO. The component values of C3 and R6 were selected so that it generates a mean output "carrier" frequency of 220 kHz. This output frequency is modulated by the original AF input signal.

Figure 9 shows how the CD4046B VCO can be converted

to a wide-range, universal, squarewave "clock" generator that covers the nominal range 0.5 Hz to 500 kHz in three switch-selected bands. This circuit can be a useful test instrument that provides a two-phase output. It can operate in either the free-running or the gated modes.

Phase-locked circuits

The basic operating principles of the CMOS phase-locked loop (PLL) IC were explained last month. You should review that

article unless you have a clear understanding of those principles before you work with the circuits presented in the remainder of this article. They will make use of all of the PLL's circuitry.

Figure 10 shows the CD4046B organized as a wide-range signal tracker. It will com-

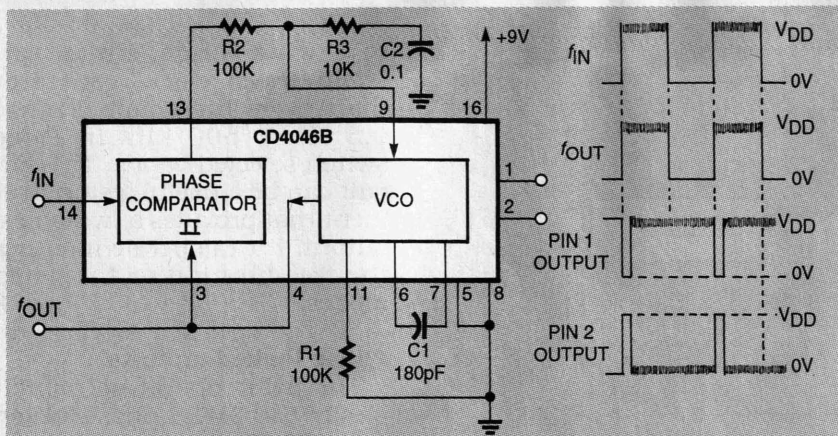


FIG. 10—WIDE-RANGE PLL SIGNAL TRACKER showing waveforms obtained when the loop is locked.

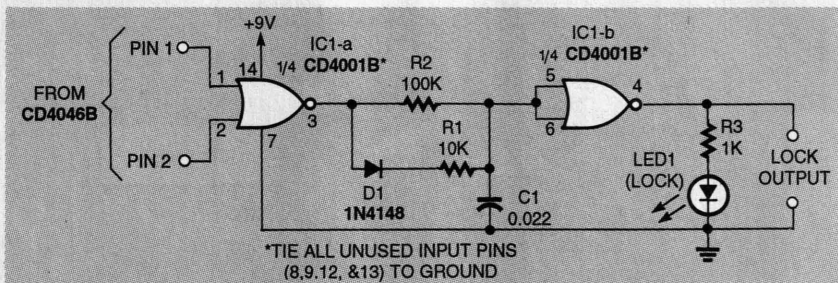


FIG. 11—PHASE-LOCKED LOOP "lock" detector/indicator.

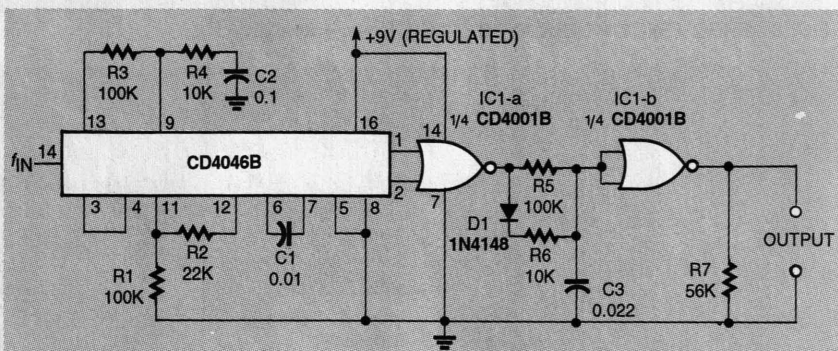


FIG. 12—PRECISION NARROW-BAND TONE SWITCH has a range of 1.8 kHz to 2.2 kHz.

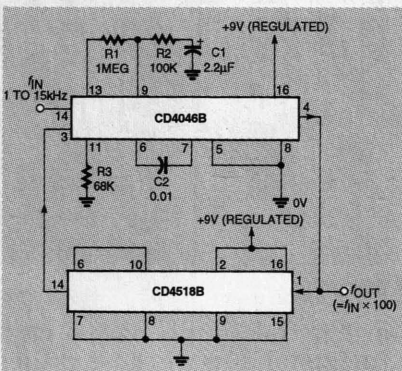


FIG. 13—A LOW FREQUENCY $\times 100$ multiplier/pre-scaler circuit.

pute and track any input signal from 100 Hz to 100 kHz, provided that the input signal to pin 14 switches fully between the zero and one logic levels.

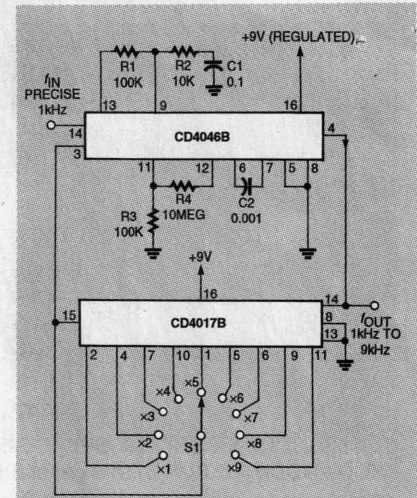


FIG. 14—FREQUENCY SYNTHESIZER covers a 1 to 9 kHz range.

to that obtained with pin 9 at the supply voltage (V_{DD}).

Figure 11 is a simple *lock detector/indicator* circuit that can be used in conjunction with the PLL circuit of Fig. 10. In the PLL, the output of each phase comparator is a series of pulses whose widths are proportional to the difference between its two input signals.

The output of phase comparator I is normally low, and that of phase comparator II is normally high, except for these pulses. When the PLL circuit is locked (see Fig. 10), the two outputs are almost perfect mirror images of each other.

In the lock detector/indicator circuit of Fig. 11, those features are implemented through two-input NOR gate IC1-a, which is driven from the outputs of the two comparators. If the loop is locked, the IC1-a output remains permanently low, thus driving IC1-b output high and turning on LED1.

If the loop is not locked, however, the IC1-a output is formed as a series of positive-going pulses that rapidly charge C1 through D1 and R1. This forces IC1-b low and holds LED1 off.

Figure 12 shows how a PLL circuit can be combined with a *lock indicator* to form a precision narrow-band tone switch. The VCO's maximum frequency is determined by R1 and C1, and the minimum frequency is determined by R1, R2, and C1.

The frequency is variable from about 1.8 kHz to 2.2 kHz

This circuit (and the others shown here) takes advantage of the internal wide-range phase comparator II. This circuit permits it to lock to any signal within the "span" range of the VCO. The filter formed by R2, R3, and C2 is organized as a sample-and-hold amplifier in this operating mode. Its component values determine the settling and tracking times of signal capture.

The VCO's operating frequency is determined by the values of R1 and C1 and the voltage on pin 9. The VCO "span" (and thus the capture and tracking range of the circuit) ranges from the VCO frequency value obtained with pin 9 at zero volts

with the component values shown in Fig. 12. The circuit can only lock to input signals within this frequency range.

Figures 13 and 14 are schematics for several practical *frequency multiplier* circuits. The circuit in Fig. 13 serves as a multiply by 100 frequency multiplier/prescaler that can change 1 Hz to 150 Hz input signals into 150 Hz to 15 kHz output signals.

The circuit in Fig. 14 is a simple frequency synthesizer. It is fed with a precise (crystal-derived) 1-kHz input signal, and its output is a whole-number multiple (in the range $\times 1$ to $\times 9$) of this signal. The CD4017B is organized as a programmable divide-by-N counter in this application. A single CD4017B can be replaced by a series of programmable decade counters to form a wide-range (10 Hz to 1 MHz) synthesizer. Ω

NIGHT VISION SCOPES

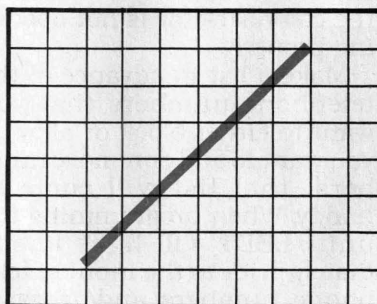
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chase a suitable IR filter at most retail camera stores, or you can stack four or five layers of completely exposed, developed film negatives between the incandescent lamp and flashlight lens. This film can be obtained as scrap from local photo developing shops. Cut four or five disks from this exposed film to fit inside the plastic or glass lens cap of your flashlight.

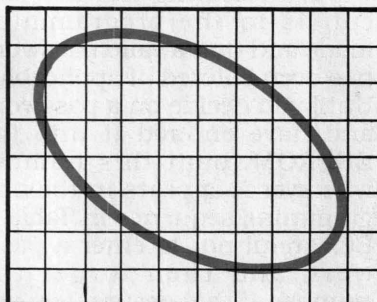
A complete kit of parts to build both of the scopes described in this article can be obtained from the source given in the parts list. If you elect to buy a surplus image tube to make a night-vision scope from scratch, purchase or obtain a "fast" camera lens and a magnifying glass for use as an eyepiece. You can then assemble all of these parts in a suitable metal or plastic tube. The power supply described here will power most imaging tubes, regardless of their size or country of origin. Ω

FUNCTION GENERATORS

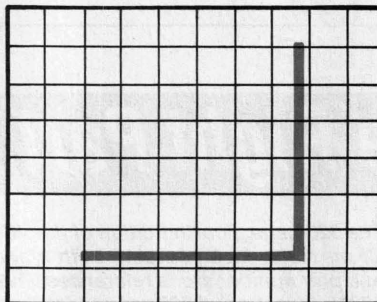
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a



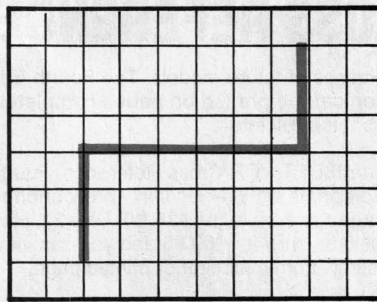
b



c



d



e

Note: A copy of B&K Precision's Guidebook to Function Generators will be sent to any reader free of charge by addressing a request in writing to Guidebook, B + K Precision, 6470 West Cortland Street, Chicago, IL 60635.

nature represents the forward voltage drop. The vertical part of the signature represents the forward current, and the horizontal part represents the reverse voltage drop.

In the waveform for the Zener diode in Fig. 9-d, the forward current is a function of the forward voltage. But when the reverse voltage equals the PN junction breakdown voltage, reverse current increases rapidly, producing a vertical line in the lower left quadrant of the screen. This line is the *break-over point* or Zener voltage, and it is established by the knee in the signature.

The signature technique can be applied to test and explain the operation of all electronic components. It is simple to use, it can speed troubleshooting, and it works well on unpowered circuit boards. Even electronic service centers operating under tight budget constraints can afford this method.

It is worth noting that two functionally identical ICs which seem to be operating normally can have different pin signatures because of differences in chip fabrication. You might encounter this when testing functionally identical IC's from different manufacturers. Different signatures do not necessarily indicate a device fault.

With experience in the careful interpretation of signatures, signature analysis can help you to identify defective components quickly—even those with marginal problems. Defective ICs (open-circuited or short-circuited) can be isolated rapidly by persons with little or no experience doing this. Ω

FIG. 9—NORMALIZED CURRENT VS. VOLTAGE SIGNATURES for electronic components: resistor (a), inductor or capacitor (b), silicon signal diode (c), light-emitting diode (d), and Zener diode (e).

DEMODULATE TV SIGNALS WITH A VIDEO IF STRIP

Separate baseband video from its carrier wave with these low-cost demodulators.

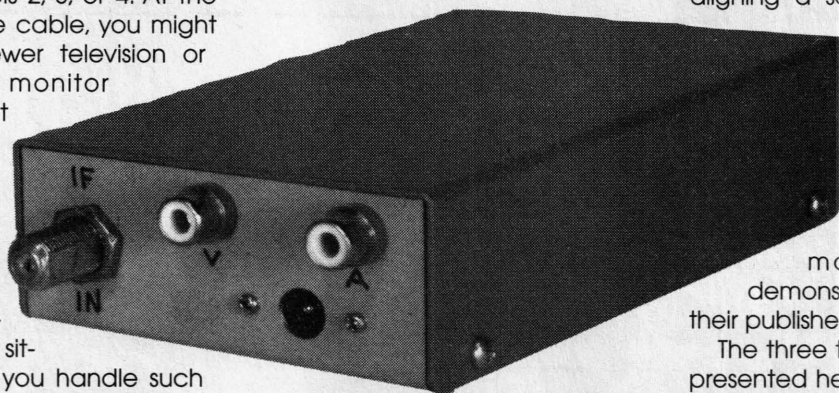
When working with video signals (the content, not the actual signal itself), it is important to know whether the source of a signal is a baseband signal or modulated on a radio-frequency (RF) carrier. For example, video sources such as video cameras and most VCRs might have a video-out jack that supplies baseband video, but older or "bottom-line" VCRs might have only a TV-antenna output—perhaps on a lower VHF frequency such as Channels 2, 3, or 4. At the other end of the cable, you might not have a newer television or home-theatre monitor that has a direct video-input jack to handle the computer video, or you might only have a baseband monitor available for the VCR situation. How do you handle such mismatches?

One device that is readily available often at low prices on the surplus market is a video modulator. That device (as its name implies) takes a baseband video signal and accompanying audio and modulates it onto a television channel. It is then a simple matter to connect the modulator's output to the antenna jack of a television and tune the set to the correct channel.

Unfortunately, if you have an RF-modulated video signal and need to convert it to baseband video the situation is not so simple. One solution might be to salvage the tuner and intermediate-frequency (IF) strip from a junked television. Assuming that those components are functional, you'll still need the proper supply voltages to run the

setup. An alternative would be to use a modern VCR, but it can be large and expensive if you need to dedicate a unit to your project. There is also the possibility of family disharmony if you keep "borrowing" the VCR if someone else wants to watch a tape or record a program for later viewing.

**WILLIAM SHEETS
AND
RUDOLF F. GRAF**



There are also situations where even those solutions will not work, such as with the partially demodulated output of a TV tuner or the reception of amateur-, satellite-, or microwave-television broadcasts. In the first situation, television audio and video signals are modulated together on a subcarrier frequency in the 44-MHz range, whereas the other situations usually use a frequency-modulated (FM) format at around 70 MHz.

In all of those cases, the best solution would be to have a dedicated IF circuit tailored to the particular needs of the received signal. Years ago, the construction of an IF circuit beyond a simple 455-kHz AM-radio system was consid-

ered to be in the realm of experienced RF engineers. Even if a good design was available, sophisticated and specialized test equipment was needed to set up and align such circuitry. That equipment was both expensive and difficult to find for the average experimenter or hobbyist.

Today, the availability of prepackaged surface-acoustic-wave (SAW) filters and related ICs have taken much of the pain and effort out of aligning a sophisticated IF circuit.

The filters have fixed bandwidth characteristics, eliminating the need for alignment. Even the basic IF circuits themselves are often provided by IC manufacturers as a demonstration application in their published data sheets.

The three types of Video IF Strips presented here will cover all of the above-mentioned situations easily: a 44-MHz demodulator for use with a TV tuner, a 66-MHz strip for separating video and audio from VHF Channel 3 or 4 signals, and a 70-MHz unit for use with FM television. Each device has a standard video output that can be connected to any NTSC-compatible monitor. They are compact, low in cost, and easy to tune and set up using only common test equipment that is readily available; in fact, the video experimenter probably already has that equipment on the workbench!

The IF66. We'll start by examining the schematic diagram of the 66-MHz unit, which will be referred to as the IF66; its schematic diagram is shown in Fig. 1. A video signal at 1 mV or greater on a 60- or 66-MHz

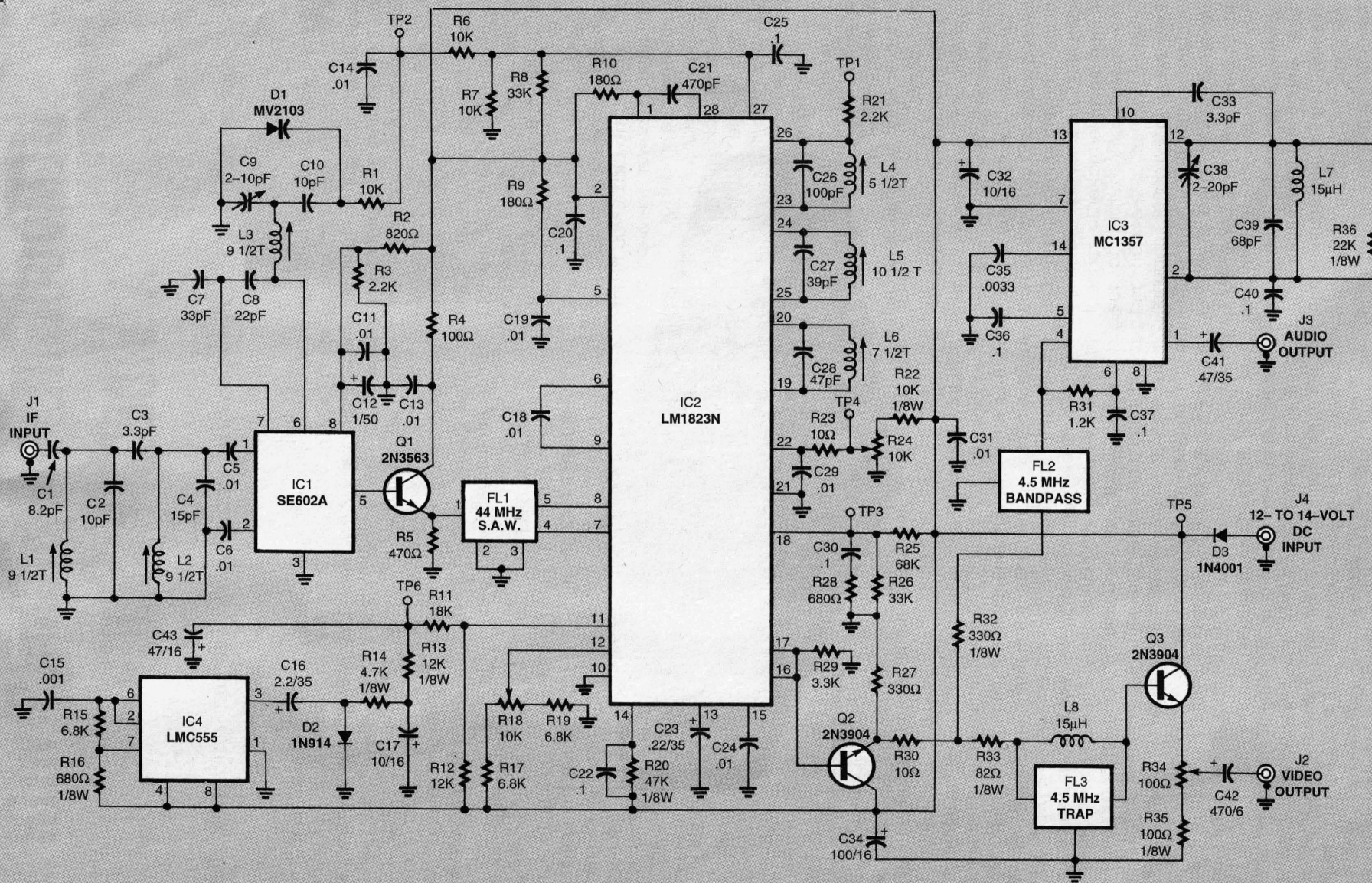


Fig. 1. The 66-MHz version of the Video IF can display any television signal on channel 3 or 4 on an NTSC video monitor. With a slight modification, the unit can be tuned to work with channels 2 through 6.

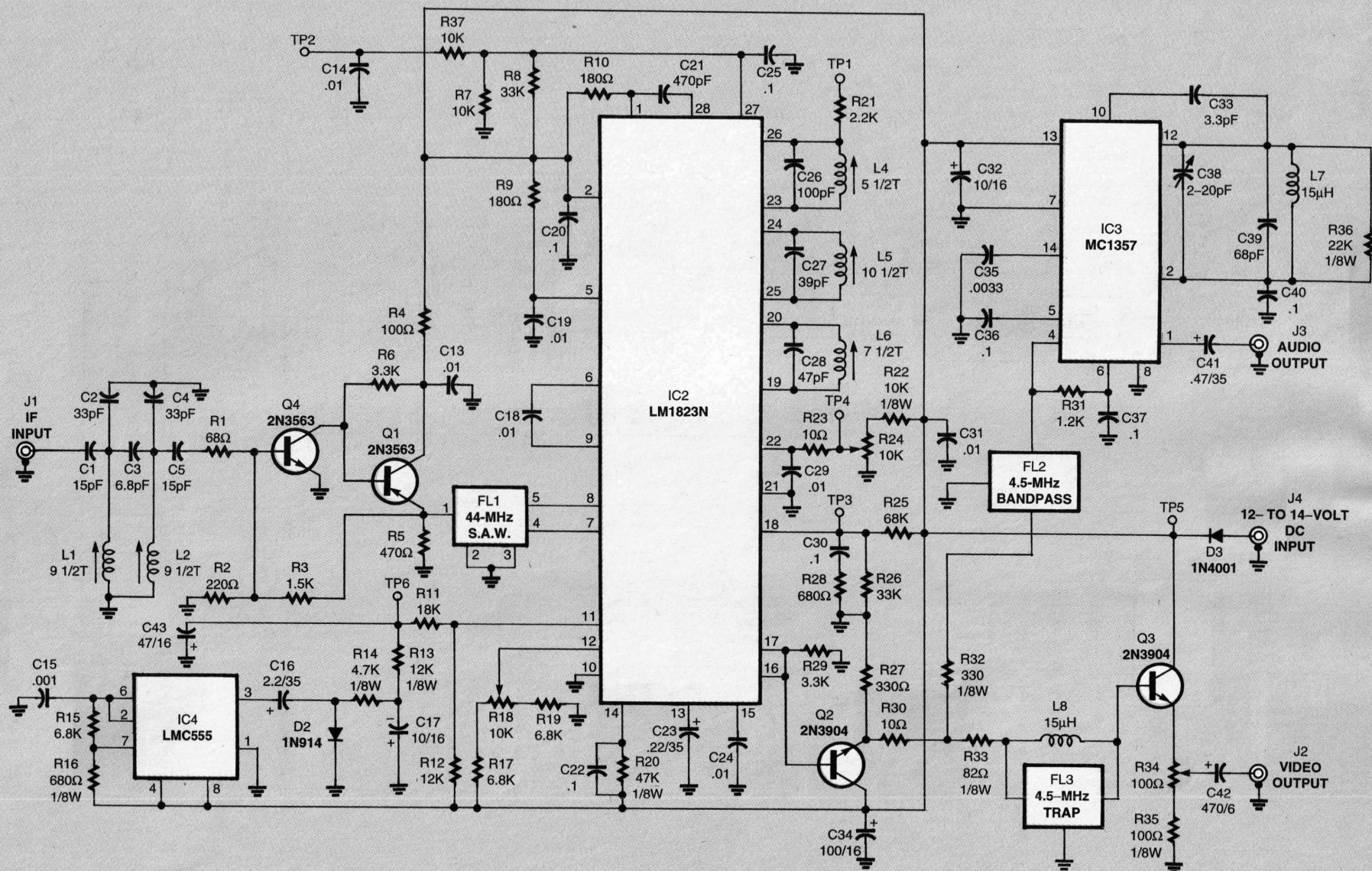


Fig. 2. The IF44 can be used to view the video from a standard TV tuner.

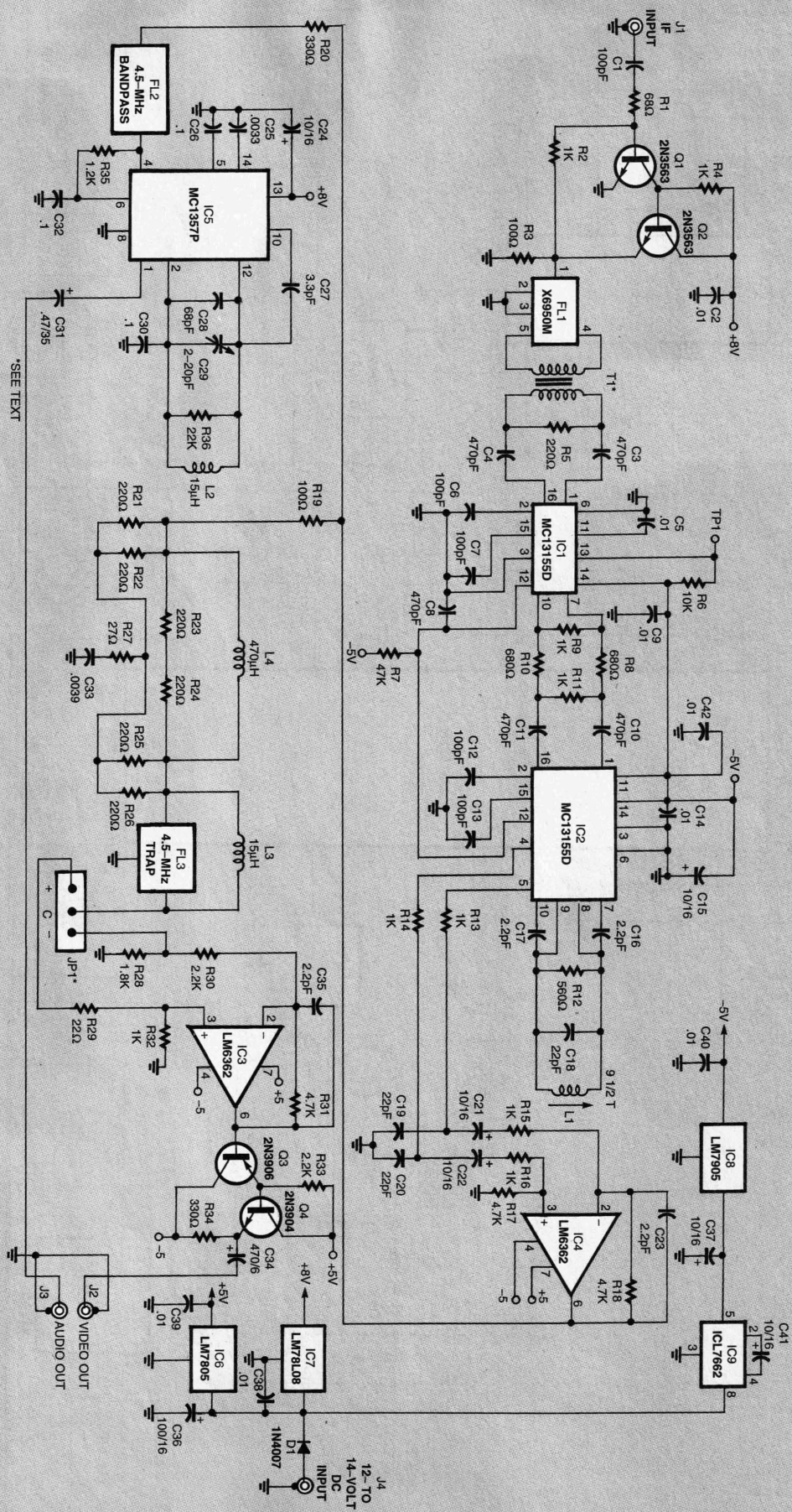


Fig. 3. This version of the Video IF can be used to view FM television signals as found on some satellite broadcasts and amateur television experiments in the microwave bands.

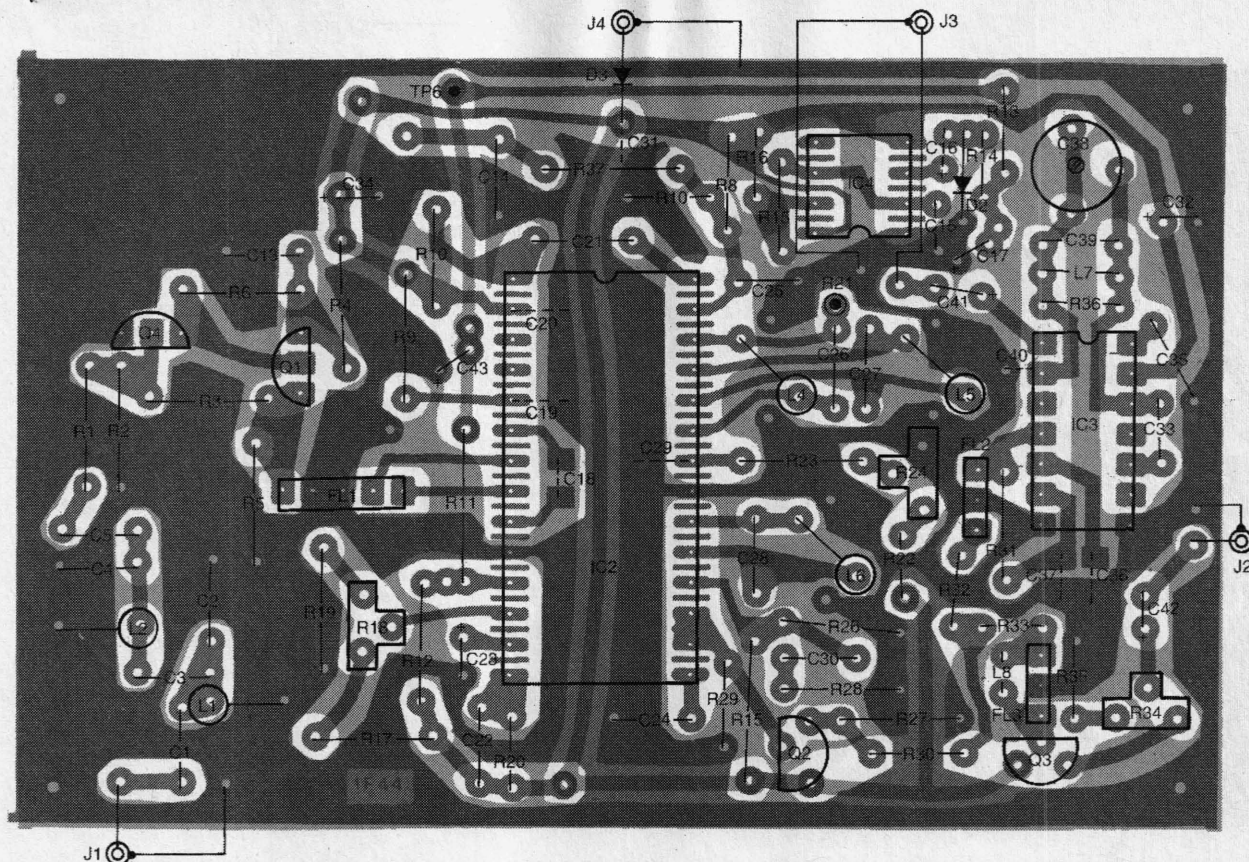


Fig. 4. The IF44 fits neatly onto a double-sided PC board. Note that a few surface-mounted capacitors are mounted on the solder side of the board.

carrier (Channel 3 or 4, respectively) is fed through J1 to a bandpass filter consisting of C1-C4, L1, and L2. That filter is designed to pass frequencies between 59 and 73 MHz at the 3-dB points. The filtered signal is fed to mixer IC1 through C5, where the incoming signal is mixed with an internally generated local-oscillator signal. The local-oscillator frequency is set by L3, C7-C10, and D1. That frequency needs to be the sum of the IF frequency (45.75 MHz) and the incoming carrier signal (61.25 or 67.25 MHz)—107 or 113 MHz. Since IC1 produces both the sum and difference frequencies, only the 45.75-MHz signal will be used. A voltage from an automatic fine-tuning circuit (AFT) is applied to D1 in order to keep the local oscillator on the needed frequency. The output signals of IC1 are fed to Q1 to provide a match between IC1's high-impedance output and the low-impedance input of FL1, a surface-acoustic-wave filter. Only the 45.75-MHz IF signal is passed through FL1 to IC2.

The heart of the unit is IC2, which contains a five-stage IF amplifier, a video detector, and the AFT circuit mentioned before. The video detector is tuned by C28 and L6, with L4 and L5 setting the operation of the limiter and AFT circuits. Note that R21 does not appear to be needed for operation of the circuit. If R21 is grounded (via TP1), the AFT circuit can be disabled if need be.

The automatic gain-control (AGC) circuitry is divided between IC4 and a portion of IC2. A square-wave between 25 and 35 kHz is generated by IC4, rectified by D2, and filtered by R14 and C17 to generate a negative-bias voltage of about -3.8 volts. The actual AGC level is created in IC2, and is set by

R18. When a sufficiently strong video signal is fed to IC2, pin 12 starts to draw current through R12, grounding that point. The 1-volt level that would normally be at TP6 drops to about -3 volts. If needed, TP6 can be used to reduce the gain of an external downconverter unit. The AGC level set by R18 must be balanced so that the IF66 does not overload on strong signals, yet will not be low enough to drop the gain too soon on any external equipment connected to TP6.

Composite video at 2 to 3 volts peak-to-peak appears at pin 17 of IC2. The signal is buffered by Q2. The video and audio portions of the composite signal are separated by FL2 (which passes only the audio subcarrier) and FL3 (which traps the audio subcarrier and passes the video signal). Removing the audio subcarrier from the video signal prevents any sound-interference patterns from disturbing the final picture. The video signal is amplified by Q3; the final voltage level is set by R34. The video signal

TABLE 1
COIL WINDING DATA

L1	9½ turns
L2	9½ turns
L3	9½ turns
L4	5½ turns
L5	10½ turns
L6	7½ turns

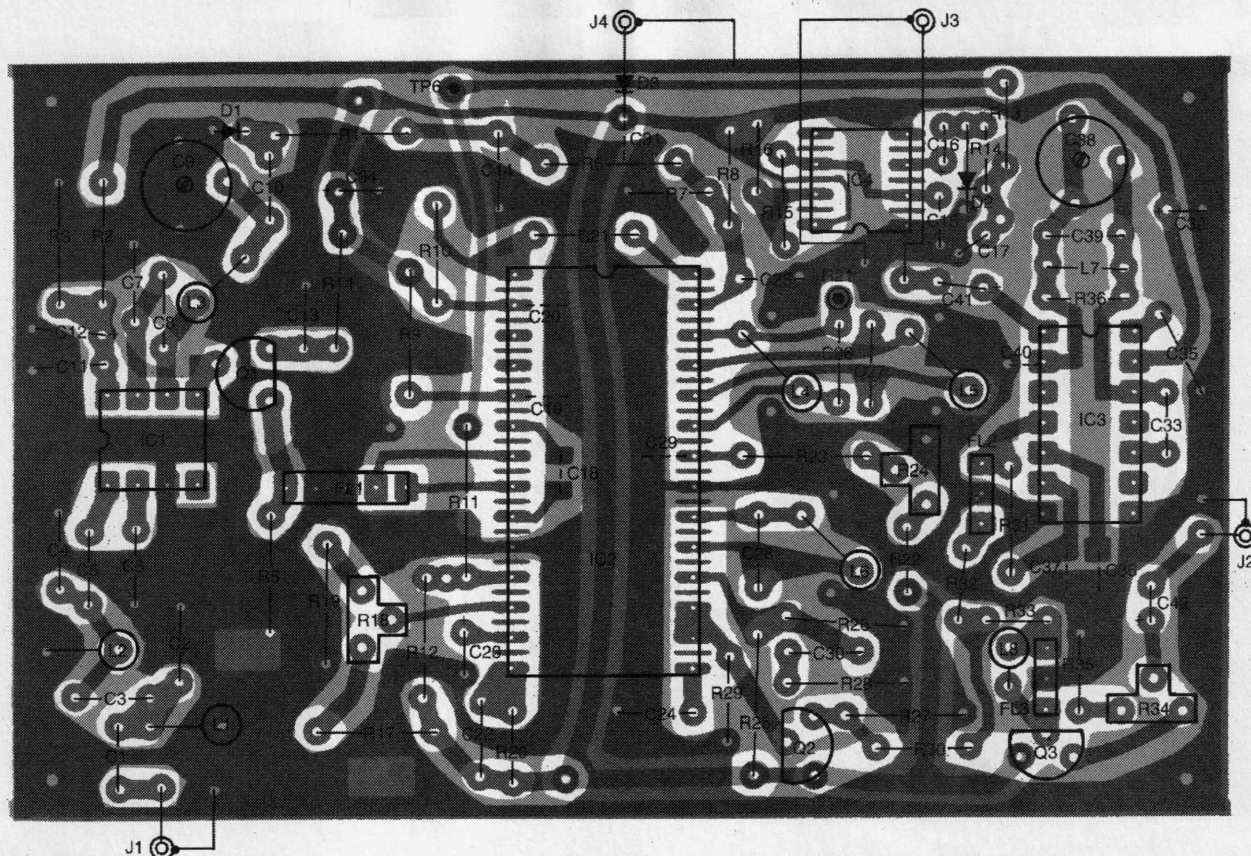


Fig. 5. Since the IF66 only differs from the IF44 in its input circuit, the layout of the two units is almost identical.

appearing on J2 is between 1- and 2-volts peak-to-peak across a 75-ohm load with a negative-going sync—a standard interface level.

The audio signal passed by FL2 is demodulated by IC3 since it is a frequency-modulated (FM) signal on a 4.5-MHz carrier. Within IC3 is an amplifier, a limiter, and a quadrature FM detector. Audio output appears at pin 1 and is passed to J3 through C41. The final audio level is 0.5 volts rms into a 5000-ohm line load. Those values work well with most audio amplifiers. Alignment of the entire audio circuit is done by C38; it is simply adjusted for the best-sounding signal.

The IF66's power supply is nominally 13.2-volts DC. At that voltage level, the current draw is about 180 mA. Protection diode D3 is included in case of an accidentally reversed supply polarity or negative transients that might damage the ICs. The supply-voltage tolerance is ± 1 volt for best performance. Do not exceed 14.4 volts, and note that less than 11 volts might cause poor performance.

TABLE 2
IF44/IF66 TEST MEASUREMENTS

D3, C34 junction	+12.6 volts
IC2 pin 2	+12.6 volts
IC1 pin 8	+7.2 volts ± 1 volt
Q1 emitter	+5.0 volts
Q3 emitter	+6.0 volts (depends on signal)
IC3 pin 13	+12.6 volts
IC3 pin 1	+6.0 volts (varies with C38)
IC4 pins 4, 8	+12.6 volts
R14, R13, C17 junction	-3.5 volts ± 0.5 volts

The IF44. The schematic diagram for the 44-MHz IF unit, which we will call the IF44, is shown in Fig. 2. A careful examination of the two schematic diagrams in Figs. 1 and 2 will quickly show that the two units are similar. In fact, they are identical with the exception of the input circuit that passes the input signal from J1 to IC2. Therefore, we will only discuss that portion of the IF44.

The IF44 is designed to work with the standard 44-MHz carrier signal supplied by standard TV-tuner hardware. A bandpass filter circuit first filters the signal from J1; it is broadly tuned to a 40- to 47-MHz window.

That frequency spread aids in rejecting any out of band frequencies that might be created by any hardware such as tuners or mixers that the IF44 is receiving its signal from. Transistors Q4 and Q1 then amplify the pre-filtered signal. Those transistors provide 23 dB of gain; R1 and R3 control the actual amount of gain. That amount of amplification will compensate for the gain loss that will come from the SAW filter. Beyond that input stage, the IF44 and IF66 are identical.

FM TV. The 70-MHz model uses a
continued on page 83

VIDEO IF

(continued from page 45)

different approach due to the nature of the video signals that it is designed to demodulate: amateur TV transmissions, experimental microwave-TV signals, satellite-TV reception, and related frequency-modulated (FM) video links...that's right, FM television! Normally, only the audio portion of a television signal is frequency-modulated; the video portion of the signal relies on amplitude modulation—basic AM. The reason for that has to do with the extreme bandwidth that FM video would require. In fact, FM-based amateur television (ATV) in the 420- to 440-MHz band is not allowed for just that reason; only AM signals are able to fit within the allotted band. However, FM TV can be used to advantage in the 902- to 928-MHz band, as well as all of the higher-frequency amateur bands. In fact, all three Video IF units described in this article are

IC3, IC4 pin 4
IC2, pins 11, 14
IC1, pin 11
IC3, IC4 pin 7
IC3, IC4 pins 2, 3, 6
IC5, pin 13
IC5, pin 1
Q2, emitter
Q1, collector
Q2, collector
Q3, emitter
Q4, emitter
IC1, IC2 pins 1, 16
IC1, IC2 pins 7, 10
IC1 pin 14 (TP1)
IC2, pins 4, 5

TABLE 3
IF70 TEST MEASUREMENTS

-5 volts ± 0.5 volts
-5 volts ± 0.5 volts
-5 volts ± 0.5 volts
+5 volts ± 0.5 volts
0.0 volts ± 0.3 volts
+8 volts ± 0.6 volts
+2 volts to +6 volts, varies with C29 setting
+0.7 volts to +1.0 volts
+1.3 volts to +2.0 volts
+8 volts ± 0.6 volts
+0.3 volts to +1.0 volts
0.0 volts ± 0.3 volts
-1.0 volts ± 0.2 volts
-1.8 volts ± 0.3 volts
-4.5 volts ± 0.5 volts (zero signal input)
-1.8 volts ± 0.3 volts, depends on input signal

designed to be connected with 440-, 900-, and 1300-MHz downconverters for the reception of ATV; the downconverter's gain can be controlled with the AGC outputs of the Video IF units.

Like AM radio, conventional television signals are susceptible to noise under moderate or weak signal conditions; the result is "snow" in the received picture. On the other hand,

having too strong of a signal can cause intermodulation-based distortions in the receiver front end; the result is a "herringbone" or patterns of wavy lines. Those interference patterns are caused by various beat frequencies and intermodulations that stem from nonlinearities in the receiver's circuits. In fact, the instantaneous transmitted power of television transmitters varies due to the nature of

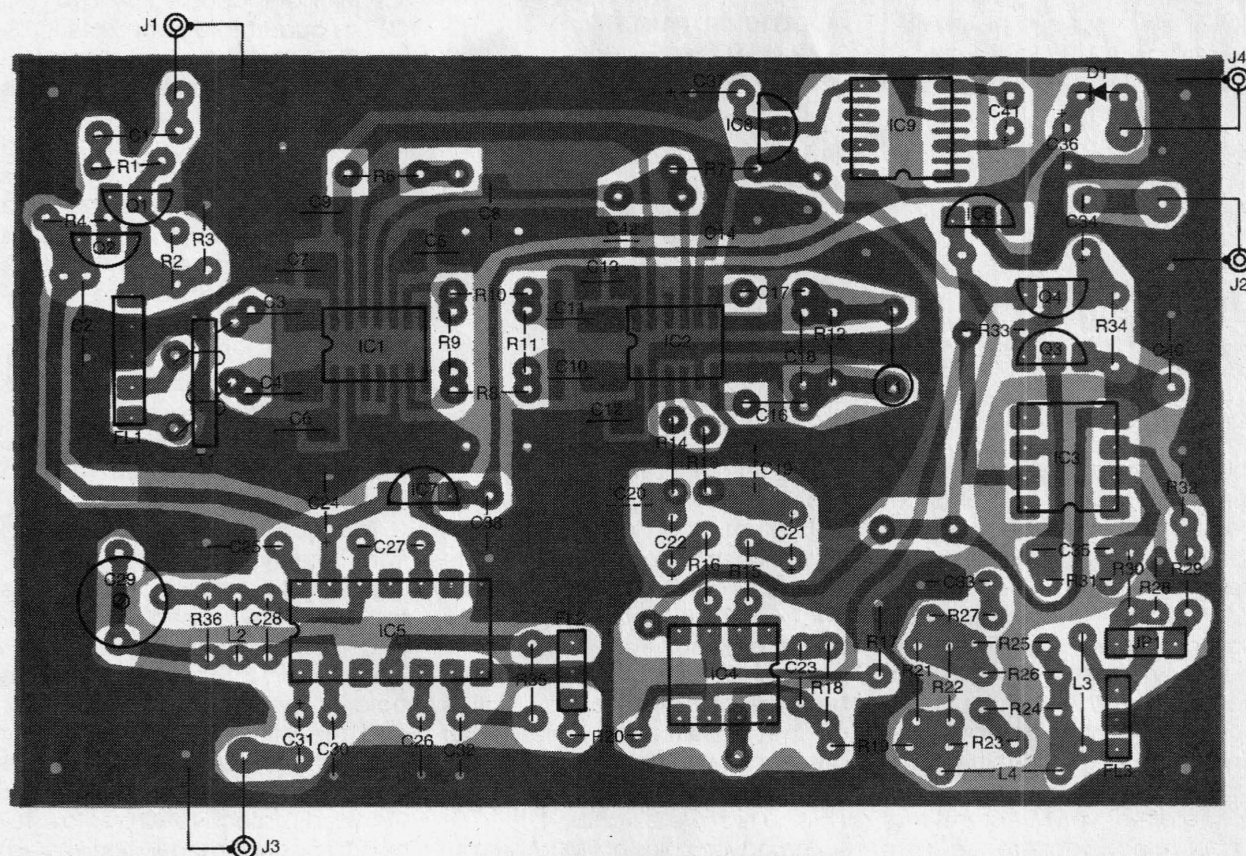


Fig. 6. The IF70 makes use of more surface-mounted components. Be careful not to cause a solder bridge when soldering IC1 or IC2 to the board—you might have to remove the chip completely in order to clear the traces underneath the component of excess solder.

PARTS LIST FOR THE 44-MHZ VIDEO IF

SEMICONDUCTORS

IC1—not used
 IC2—LM1823N video IF detector, integrated circuit
 IC3—MC1357 audio IF detector, integrated circuit
 IC4—LMC555 timer, integrated circuit
 D1—not used
 D2—1N914 silicon diode
 D3—1N4001 silicon diode
 Q1, Q4—2N3563 NPN transistor
 Q2, Q3—2N3904 NPN transistor

RESISTORS

(All resistors are 1/4-watt, 5% units unless otherwise noted.)

R1—68-ohm
 R2—220-ohm
 R3—1500-ohm
 R4—100-ohm
 R5—470-ohm
 R6, R29—3300-ohm
 R7, R37—10,000-ohm
 R8, R26—33,000-ohm
 R9, R10—180-ohm
 R11—18,000-ohm
 R12—12,000-ohm
 R13—12,000-ohm, 1/8-watt
 R14—4700-ohm, 1/8-watt
 R15, R17, R19—6800-ohm
 R16—680-ohm, 1/8-watt
 R18, R24—10,000-ohm potentiometer, PC-mount
 R20—47,000-ohm, 1/8-watt
 R21—2200-ohm
 R22—10,000-ohm, 1/8-watt
 R23, R30—10-ohm
 R25—68,000-ohm
 R27—330-ohm
 R28—680-ohm
 R31—1200-ohm
 R32—330-ohm, 1/8-watt
 R33—82-ohm, 1/8-watt
 R34—100-ohm potentiometer, PC-mount
 R35—100-ohm, 1/8-watt

R36—22,000-ohm, 1/8-watt

CAPACITORS

C1, C5—15-pF, ceramic-disc
 C2, C4—33-pF, ceramic-disc
 C3—6.8-pF, ceramic-disc
 C6—C12—not used
 C13, C14, C24—0.01-μF, ceramic-disc
 C15—0.001-μF, Mylar
 C16—2.2-μF, 35-WVDC, electrolytic
 C17, C32—10-μF, 16-WVDC, electrolytic
 C18, C19, C29, C31—0.01-mF, ceramic, surface-mount
 C20, C36, C37, C40—0.1-μF, ceramic, surface-mount
 C21—470-pF, ceramic-disc
 C22, C25, C30—0.1-μF, Mylar
 C23—0.22-μF, 35-WVDC, electrolytic
 C26—100-pF, ceramic-disc
 C27—39-pF, ceramic-disc
 C28—47-pF, ceramic-disc
 C33—3.3-pF, ceramic-disc
 C34—100μF, 16-WVDC, electrolytic
 C35—0.0033-μF, Mylar
 C38—2–20-pF, ceramic trimmer
 C39—68-pF, ceramic-disc
 C41—0.47-μF, 35-WVDC, electrolytic
 C42—470-μF, 6-WVDC, electrolytic
 C43—47-μF, 16-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

FL1—44-MHz surface-acoustic-wave filter (Murata)
 FL2—4.5-MHz bandpass filter (Murata)
 FL3—4.5-MHz trap filter (Murata)
 J1—F-style connector, panel-mount
 J2, J3—RCA-style jack, panel-mount
 J4—Co-axial power jack
 L1—L6—22-gauge magnet wire wound on Cambion blue slugs (see text)
 L7, L8—15-μH RF choke
 12–14-volt DC wall-mounted transformer (RadioShack 22-504 or similar, see text), wire, hardware, etc.

amplitude-modulated video—the transmitted power changes depending on whether any particular portion of the video picture is lighter or darker.

FM television offers certain advantages over AM TV in the same way that FM radio has advantages over AM radio, including better signal-to-noise ratios, the ability to run the transmitter at a steady power output, and the "capture effect" in FM systems that tend to reject interference from weaker signals on the same frequency. However, AM does have the advantage when it comes to

multipath-like interference such as "ghosting"; while annoying in an AM system, it can totally destroy FM reception because of the severe distortion of the received video. Since FM TV is used mostly at UHF and microwave frequencies, multipath distortion is all but eliminated because of the directional nature of the transmission; unless your antenna is pointing directly at the transmitter, it is difficult, if not impossible, to receive a signal.

The advantages of FM TV are put to good use in satellite broadcasts. The ability of an FM receiver to "lock" onto a weak signal is perfect for

satellites; the transmitter power available in a satellite is extremely limited. **The IF70.** The schematic diagram for the IF70 is shown in Fig. 3. Like the two previously-described units, the IF70 uses standard IC demodulators and surface-acoustic-wave filters to eliminate any tricky alignment and setup adjustments.

A 70-MHz input signal at J1 is applied to preamplifier stage Q1 and Q2. The gain is set by R1 and R2 to compensate for losses that occur in the SAW filter, FL1. Transformer T1 and R5 provide a balanced drive to IC1, an IF amplifier circuit that can handle FM signals as high as 300 MHz. A DC level between 0 to 3 volts appears across R6; it has a fairly logarithmic response at about one volt per 20 dB of increase in signal strength. If needed, that signal can be used as a received signal-strength indicator (RSSI).

Whereas IC1 is used as an amplifier, IC2 is set up as a limiter and quadrature detector to recover the video information. The output of IC2's limiter is coupled by C16 and C17 to quadrature network L1, C18, and R12. The recovered video-output signal, in a differential format, is available at pins 4 and 5 of IC2.

The differential video is amplified by IC4. That amplifier, configured as a differential amplifier, produces a single-ended output. The video is then de-emphasized by R21–R27, C33, and L4. The purpose of de-emphasis is to correct the frequency distortion that is deliberately introduced in the transmitted video in order to improve the signal-to-noise ratio. That method is similar to the audio pre-emphasis method that is used in FM radio.

The sound subcarrier is removed from the video signal by FL3; removing it helps reduce distortions that might be caused by a sound-to-color beat-interference.

Depending on modulation polarity, downconverter circuitry, and the effect of any frequency mixing, the output signal can have a negative or positive sync polarity. Generally, black-to-white low-to-high frequency modulation is used, but some receivers can change that. In any case, some means of polarity selection is needed. The

de-emphasized video is fed to either the inverting or non-inverting input of IC3 by means of a jumper block on JP1. If you find that you need to change the video polarity frequently, JP1 can be replaced by a single-pole, double-throw switch.

The output of IC3 feeds a video buffer made from Q3 and Q4, giving the IF70 the ability to drive a 75-ohm load connected to J2.

The audio subcarrier is separated from the composite signal by FL2. The audio is then amplified, limited, and detected by a conventional quadrature-detector circuit; IC5 has been specifically designed for that purpose. The center frequency and bandwidth is set by C28, C29, L2, and R36. The adjustment of C29 sets the center frequency of the audio bandwidth. The audio output is around 0.5-volts rms into a 5000-ohm load. That amount of output drive works well with most audio gear.

Regulators IC6, IC7, and IC8 supply the +5-, +8-, and -5-volt DC levels that are needed to power the IF70. The voltage supplied to the regulators must be at least 10 volts so that IC7 has at least a 2-volt differential between its input and output pins; anything lower than that will cause the regulator to shut down. Taking the voltage drop of D1 into consideration, the supply voltage connected to J4 should be at least 11 volts; 12 volts would be ideal.

A -5-volt supply is needed for IC1 and IC2 in order to have effective RF grounding without any elaborate RF decoupling and bypassing. That voltage is also needed for op-amps IC3 and IC4. A charge-pump DC-DC converter, IC9, produces a negative voltage that is regulated by IC8. That method of obtaining a negative voltage works well because only about 35 mA is needed.

Construction. Because of the high frequencies involved, the various Video IF units are best built on PC boards. Foil patterns for the double-sided PC boards have been included here if you wish to etch your own board. As an alternative, a pre-etched PC board is available as part of a complete kit; it is not available separately. For more information on the kits for all three units,

PARTS LIST FOR THE 66-MHZ VIDEO IF

SEMICONDUCTORS

IC1—SE602A video demodulator, integrated circuit
IC2—LM1823N video IF detector, integrated circuit
IC3—MC1357 audio IF detector, integrated circuit
IC4—LMC555 timer, integrated circuit
D1—MV2103 Varactor diode
D2—1N914 silicon diode
D3—1N4001 silicon diode
Q1—2N3563 NPN transistor
Q2, Q3—2N3904 NPN transistor

RESISTORS

(All resistors are 1/4-watt, 5% units unless otherwise noted.)

R1, R6, R7—10,000-ohm
R2—820-ohm
R3, R21—2200-ohm
R4—100-ohm
R5—470-ohm
R8, R26—33,000-ohm
R9, R10—180-ohm
R11—18,000-ohm
R12—12,000-ohm
R13—12,000-ohm, 1/2-watt
R14—4700-ohm, 1/2-watt
R15, R17, R19—6800-ohm
R16—680-ohm, 1/2-watt
R18, R24—10,000-ohm potentiometer, PC-mount
R20—47,000-ohm, 1/2-watt
R22—10,000-ohm, 1/2-watt
R23, R30—10-ohm
R25—68,000-ohm
R27—330-ohm
R28—680-ohm
R29—3300-ohm
R31—1200-ohm
R32—330-ohm, 1/2-watt
R33—82-ohm, 1/2-watt
R34—100-ohm potentiometer, PC-mount
R35—100-ohm, 1/2-watt
R36—22,000-ohm, 1/2-watt

CAPACITORS

C1—8.2-pF, ceramic-disc
C2, C10—10-pF, ceramic-disc
C3, C33—3.3-pF, ceramic-disc
C4—15-pF, ceramic-disc
C5, C6, C11, C13, C14, C24—0.01-μF, ceramic-disc
C7—33-pF, ceramic-disc
C8—22-pF, ceramic-disc
C9—2-10-pF, trimmer
C12—1-μF, 50-WVDC, electrolytic
C15—0.001-μF, Mylar
C16—2.2-μF, 35-WVDC, electrolytic
C17, C32—10-μF, 16-WVDC, electrolytic
C18, C19, C29, C31—0.01-μF, surface-mount
C20, C36, C37, C40—0.1-μF, surface-mount
C21—470-pF, ceramic-disc
C22, C25, C30—0.1-μF, Mylar
C23—0.22-μF, 35-WVDC, electrolytic
C26—100-pF, ceramic-disc
C27—39-pF, ceramic-disc
C28—47-pF, ceramic-disc
C34—100-μF, 16-WVDC, electrolytic
C35—0.0033-μF, Mylar
C38—2-20-pF, trimmer
C39—68-pF, ceramic-disc
C41—0.47-μF, 35-WVDC, electrolytic
C42—470-μF, 6-WVDC, electrolytic
C43—47-μF, 16-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

FL1—44-MHz surface-acoustic-wave filter (Murata)
FL2—4.5-MHz bandpass filter (Murata)
FL3—4.5-MHz trap filter (Murata)
J1—F-style connector, panel-mount
J2, J3—RCA-style jack, panel-mount
J4—Co-axial power jack
L1-L6—22-gauge magnet wire wound on Cambion blue slug (see text)
L7, L8—15-μH RF choke
12-14-volt DC wall-mounted transformer (RadioShack 22-504 or simi-

see the Parts List for the IF70.

The boards for the three different versions of the Video IF are all the same size. If you are going to etch your own board, be sure to match up the correct pair of foil patterns—especially if you will be building the 44-MHz or 66-MHz version!

In building a board that is as tightly packed as the Video IF boards, it is helpful to first install a few larger parts such as trimmer caps and potentiometers; they will serve as landmarks. IC sockets may be used if desired, but only the low-profile type. As with any RF-based project,

lead lengths should be kept as short as possible. Excess lead length and socket capacitance might cause instability and other problems such as video ringing, glitches, or oscillation; the result, of course, is poor performance.

There are also several components as well as "via" holes that must have solder connections on both sides of the board. Even if the connection is a ground, it is important to make both connections. Doing that is essential for proper RF grounding.

The parts-placement diagram for

the IF44 is shown in Fig. 4. If you are building the IF66, follow Fig. 5 instead. For the IF70, use Fig. 6. Begin by inserting all of the resistors. Note that on the IF44 and IF66, R21 is mounted vertically with the unmounted end left free. That end of R21 forms a test point; it can be formed into a small loop of wire as close to the body of the resistor as possible.

If IC sockets are to be used, install them next. When installing C9 on the IF66, be careful not to melt the plastic body when making the solder connections on the top side of the PC board. When installing the through-hole capacitors, any preformed leads should be straightened with pliers so that the component sits as close to the PC board as possible. Watch polarity of the electrolytic capacitors. DO NOT install the chip capacitors at this time; they will be mounted last.

Next, mount the transistors, diodes, and potentiometers. All of the transistors should be $\frac{1}{8}$ inch from surface of the board. Once again, don't forget to solder any leads that have pads on both sides of the PC board. On the IF44 and IF66, D3 is installed vertically with the cathode (banded end) toward the PC board. The anode lead is formed into a terminal that will be connected to J4 later. When installing the SAW filters, double-check the orientation of the parts before soldering them.

Form the coils by winding magnet wire around a screw; Fig. 7 shows the details with Table 1 giving the number of turns needed for the various coils. Once they are made, install them where needed. Note that not all coils are used in any particular unit. On the IF70, a toroidal transformer will also be needed; its construction details are given in Fig. 8.

The surface-mount capacitors are installed on the solder side of the board; their positions are shown with a dotted line in the parts-placement diagrams. Once the surface-mount components are installed, avoid flexing the PC board; the surface-mount components might crack or become intermittent. The surface-mount ICs on the IF70 are installed on the top side of the PC board. Be very careful to avoid sol-

PARTS LIST FOR THE 70-MHZ VIDEO IF

SEMICONDUCTORS

IC1, IC2—MC13155D FM demodulator, integrated circuit
IC3, IC4—LM6362 high-speed op-amp, integrated circuit
IC5—MC1357P audio IF detector, integrated circuit
IC6—LM7805 or LM78L05 5-volt voltage regulator, integrated circuit
IC7—LM78L08 8-volt voltage regulator, integrated circuit
IC8—LM7905 or LM79L05 -5-volt voltage regulator, integrated circuit
IC9—ICL7662 voltage converter, integrated circuit
D1—1N4007 silicon diode
Q1, Q2—2N3563 NPN transistor
Q3—2N3906 NPN transistor
Q4—2N3904 PNP transistor

RESISTORS

(All resistors are $\frac{1}{8}$ -watt, 5% units.)

R1—68-ohm
R2, R4, R9, R11, R13—R16, R32—1000-ohm
R3, R19—100-ohm
R5, R21—R26—220-ohm
R6—10,000-ohm
R7—47,000-ohm
R8, R10—680-ohm
R12—560-ohm
R17, R18, R31—4700-ohm
R20, R34—330-ohm
R27—27-ohm
R28—1800-ohm
R29—22-ohm
R30, R33—2200-ohm
R35—1200-ohm
R36—22,000-ohm

CAPACITORS

C1—100-pF, ceramic-disc
C2, C38—C40—0.01- μ F, ceramic-disc
C3, C4, C8, C10, C11—470-pF, surface-mount
C5, C9, C14, C42—0.01, μ F, surface-mount
C6, C7, C12, C13—100-pF, surface-mount

C15, C21, C22, C24, C37, C41—10- μ F, 16-WVDC, electrolytic
C16, C17, C23, C35—2.2-pF, ceramic-disc
C18—22-pF, ceramic-disc
C19, C20—22-pF, surface-mount
C25—0.0033- μ F, Mylar
C26, C30, C32—0.1- μ F, Mylar
C27—3.3-pF, ceramic-disc
C28—68-pF, ceramic-disc
C29—2-20-pF trimmer
C31—0.47- μ F, 35-WVDC, electrolytic
C33—0.0039- μ F, ceramic-disc, 5%
C34—470- μ F, 6-WVDC, electrolytic
C36—100- μ F, 16-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

FL1—70-MHz surface-acoustic-wave filter, 11-MHz bandwidth (Siemens X6950M)
FL2—4.5-MHz ceramic bandpass filter (Murata)
FL3—4.5-MHz ceramic trap filter (Murata)
JP1—3-pin header, 0.1-inch center spacing
L1—22-gauge magnet wire wound on Cambion blue slug (see text)
L2, L3—15- μ H RF choke
L4—470- μ H RF choke
T1—Toroid transformer (see text)
12-14-volt DC wall-mounted transformer (RadioShack 22-504 or similar, see text), jumper block, wire, hardware, etc.

Note: A complete kits of all board-mounted components and a drilled and etched PC board are available from North Country Radio, PO Box 53, Wykagyl Station, New Rochelle, N.Y. 10804-0053; Web: <http://www.northcountryradio.com>; IF44, \$54.75; IF66, \$ 59.75; IF70, \$ 69.75. Please include \$4.50 for shipping and handling within the US, or \$10.00 for shipments outside of the US. NY residents must add appropriate sales tax. Please note that cases, switches, power supplies, or external connectors are not

der bridges between the terminals. Repairing a solder bridge on a surface-mount IC is very difficult; often, the IC must be removed to get at the underlying trace.

Install a short lengths of stiff wire with a loop at the end at TP6; if you do the same for the IF-input, audio-output, and video-output terminals, it will make wiring the PC board into the case easier. Install the through-hole ICs in the board by either inserting them into the sockets or soldering them directly to the

board. Keep in mind that the ICs are static sensitive.

Carefully inspect your work for solder shorts, poor joints, missing parts, incorrect parts placement, and improper orientation of polarized parts such as ICs, filters, transistors, and electrolytic capacitors. Once you are satisfied with your work, the Video IF can be mounted in a suitable metal or plastic case with input, output, and power connectors of your choice. A metal box is preferred since it provides shielding

and reduces noise pickup. Recommended connectors for the Video IF are F or BNC connectors for J1, RCA connectors for J2 and J3, and a co-axial power jack for J4; the choice for J4 will depend on your available power supply.

As the circuitry uses about 2 to 3 watts of DC power, it will run somewhat warm and should be ventilated if it will be used in a small space. The Video IF works best with a high-

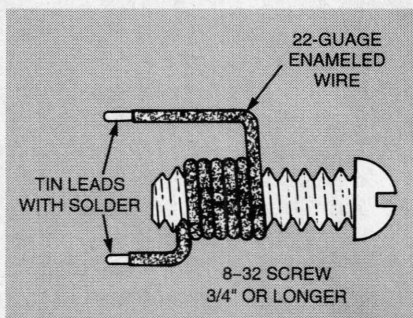


Fig. 7. The coils needed by the Video IF units are easy to make when a machine screw is used as a winding form.

quality wall-mounted 12-volt power supply that is capable of at least 200 mA. You should avoid using inexpensive units as they are often poorly filtered and unregulated; they can blow out the ICs due to overvoltage.

When everything is assembled, you are ready to test the unit.

Testing the IF66. Testing and calibrating the IF66 is straightforward and only requires simple test equipment: a digital voltmeter, a source of NTSC television signal on Channel 3 or 4 (such as the output of a VCR), and an NTSC video monitor to view the output. A computer monitor will not work unless it can display NTSC composite video. Many newer TV sets have an external video-input jack; such a set will work well. Of course, appropriate cables are also needed to connect the equipment together.

A frequency counter that is reliable to at least 200 MHz is a great help in setting up the local oscillator, although an FM-broadcast receiver covering the 88- to 108-MHz band can serve if a frequency counter is unavailable.

To begin, connect the power supply to J4. Note that D3 will prevent

any damage if the power supply polarity is accidentally reversed. Measure the current being drawn by the unit; it should be around 150 to 200 mA. If it is higher, check for possible shorts. If it is lower than 150 mA, something might be open or missing from the circuit. Make sure that all of the connections to the top side of the board—especially connections to the power-supply traces—have been installed.

Referring to the schematic diagram and the parts-placement diagram, check for the voltages listed in Table 2 with respect to ground.

Note that nothing should be getting hot. Although IC2 will run somewhat warm, it should not be uncomfortable to hold your finger on it. Any major variations from the expected readings should be investigated before proceeding. Check for parts placement, component values, solder shorts, and poor joints. Make sure that your test equipment is set up properly.

Set R18, R24, and R34 to the center of their rotation range, C9 and C38 so that their plates are $\frac{1}{4}$ meshed, and the slugs in all of the coils so that they are flush with the top of their windings. Back the slugs in L4, L5, and L6 out $2\frac{1}{2}$ turns; they should be $\frac{1}{16}$ to $\frac{3}{32}$ inch above the top of the windings.

Turn the unit off. Connect the outputs at J2 and J3 to a monitor and audio amplifier, respectively, and J1 to a video source. Temporarily connect a clip lead between TP1 (the free end of R21) and ground. With the signal source, monitor, and audio turned on, power up the Video IF. You should see at least some noise or other indication that the Video IF is working. Some audio activity should also be heard, if only an increase in noise. If not, check all connections.

Couple a frequency counter to L3 using a 2- or 3-turn loop of wire. Rotate C9 to obtain a reading of 107 MHz (for Channel 3) or 113 MHz

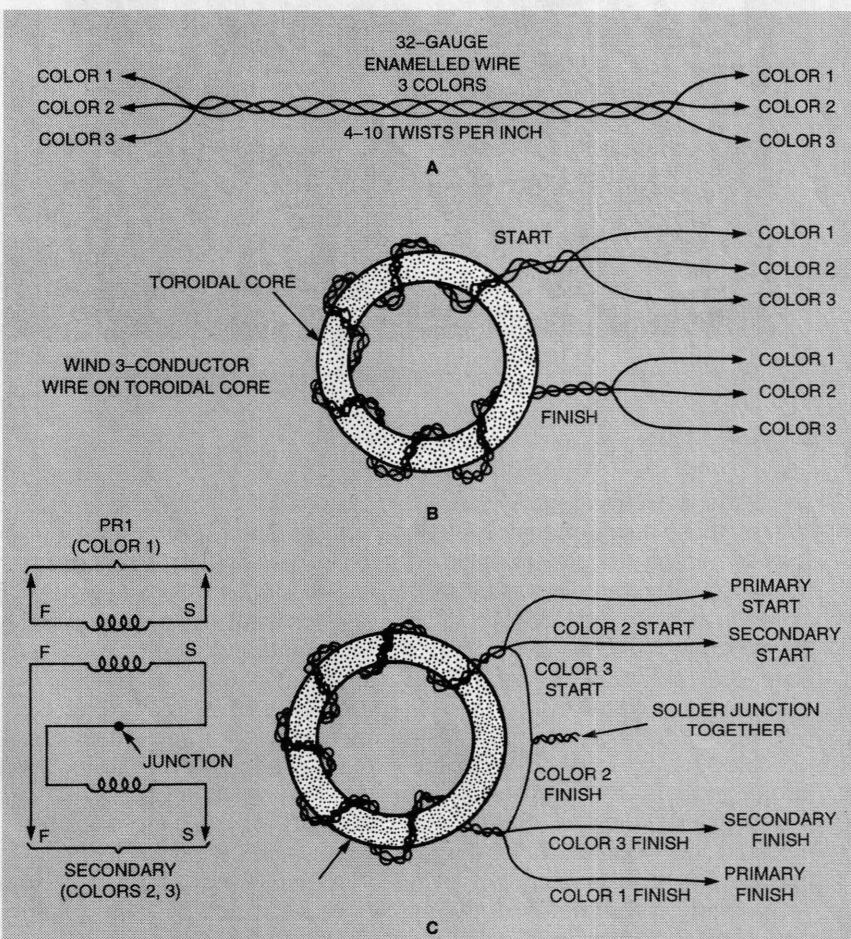
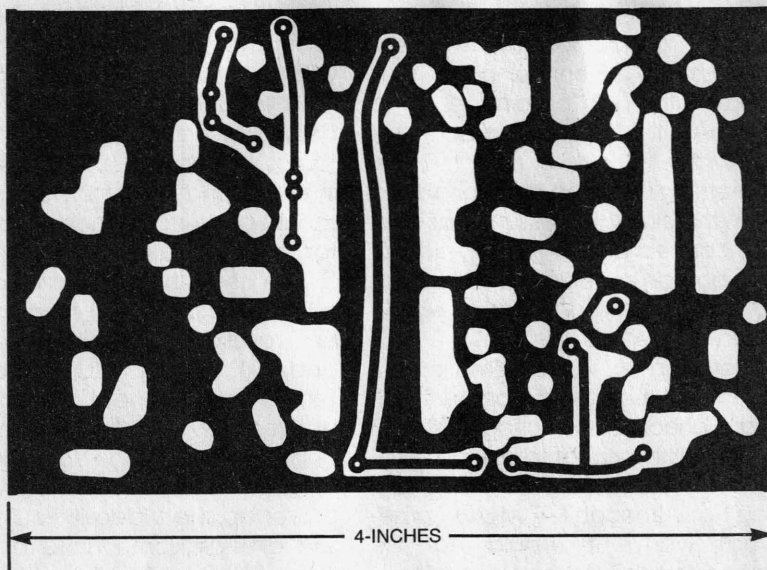
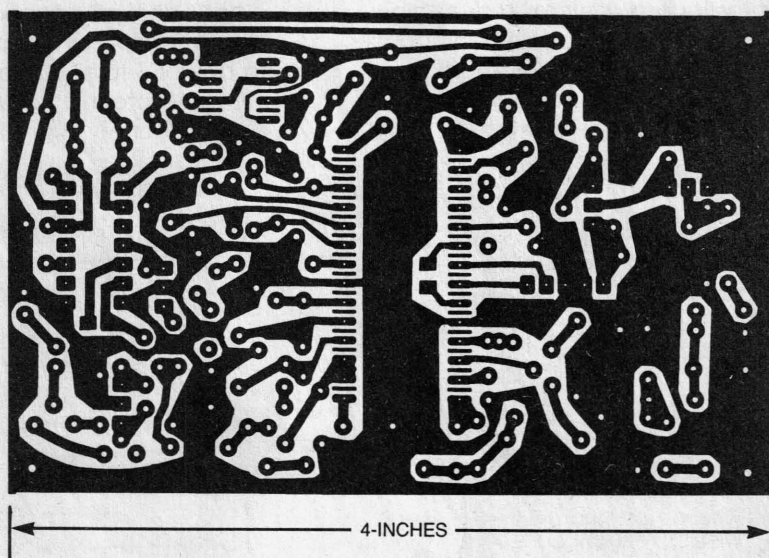


Fig. 8. Although the transformer for the IF70 is a bit more complicated to make, it is still easy to do by following these simple steps.



Here is the foil pattern for the component side of the IF44.



Here is the foil pattern for the solder side of the IF44.

(for Channel 4). Readings within 200 kHz of those frequencies are OK; they should be steady. Any reading that is jumping wildly or way off indicates a problem with the circuit. An FM receiver tuned to a quiet spot near 106 MHz can be used if no counter is available or if a reliable counter reading cannot be obtained. Listen for the local-oscillator signal as C9 is rotated. It will sound like a sudden quieting of background noise as C9 is tuned through the receiver frequency.

Connect a voltmeter between pin 18 of IC2 and ground. Adjust the slug of L6 until a video image appears on the monitor. Once a

picture is available, set L6 for a 4.5-volt reading. Connect the meter to TP4 (wiper of R24). Adjust for a reading of about 4.5 to 5 volts. Adjust R34 for the best picture quality on the monitor.

Remove the clip lead from TP1 (free end of R21). The picture may become garbled—that is OK. Connect the voltmeter to TP2 (the junction of R1, R6, and C14). Adjust the slug in L4 to bring back the picture on the monitor. The best picture quality should occur at around a 3-volt reading. That adjustment will be somewhat "sharp." You may have to retouch C9 so that the best picture is obtained at a reading of

3 to 3.5 volts at TP2. That is the AFT adjustment.

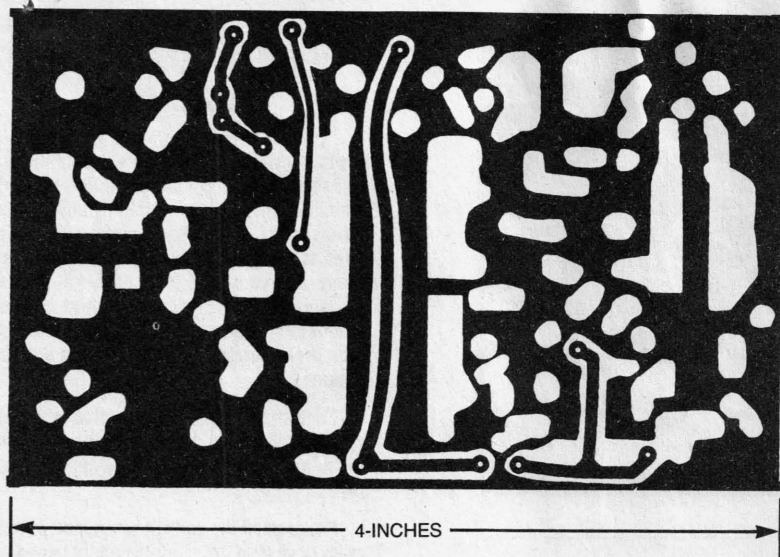
Adjust C38 for the clearest audio. If you run out of adjustment range or find that C38 is fully meshed or unmeshed at the best audio setting, change C39 to 82 pF or 56 pF, respectively. That will likely be unnecessary in most cases. Before doing that, check that the value of C33 is correct (3.3 pF, not 33 pF), L7 has continuity, and that C40 is properly installed. If there is still no audio, check that FL1 and FL2 have not been mixed up by accident and that IC3 and its associated components are correctly installed.

Try adjusting L5 for the maximum RF voltage at pin 25 of IC2. If you don't have the equipment to do that, simply experiment with the settings of L5 for the best appearing picture. That adjustment is somewhat broad; if it seems to have little effect, simply return the slug to its initial preset position (backed out 2.5 turns from flush with top).

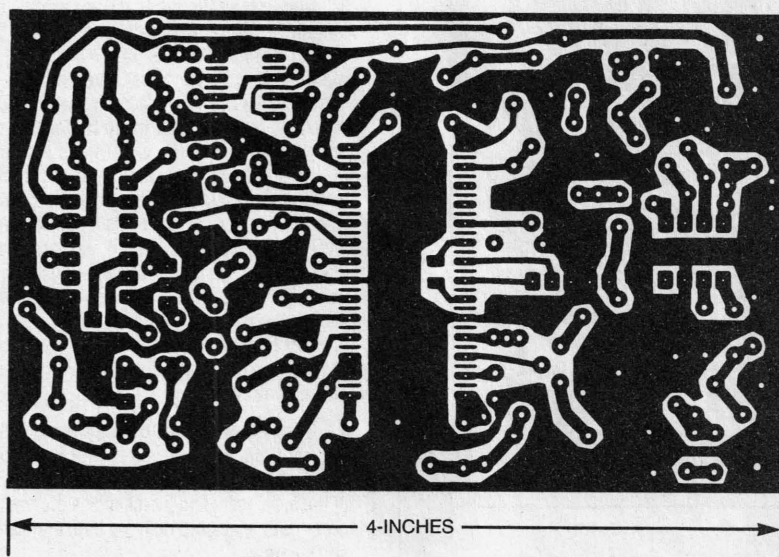
Adjust L1 and L2 for a frequency response that is within 2 dB over the range of 60–72 MHz. If no suitable test equipment is available, try adjusting the slugs for the best picture. Do not exceed 2 turns in either direction from the initial settings. If little or no improvement can be seen, return the slugs to the initial settings.

The IF66 is now considered calibrated and working. As an additional test, vary the supply voltage from 11.5 to 14.4 volts. Under no circumstances should the supply voltage exceed 14.4 volts. The Video IF should work over that range. If it does not, recheck all of the settings and readjust as needed. Sometimes you might find that L6 will have to be reset slightly.

With a slight modification, the IF66 can be adjusted to operate on Channel 2. Simply add a turn to L1, L2, and L3. The local-oscillator frequency will then be 101 MHz. If you'd like to raise the output to Channel 5 or 6, remove a turn from those coils. The local-oscillator frequency would then become 119 MHz for channel 5 and 129 MHz for channel 6. No changes would need to be made to L4, L5, or L6; the alignment procedure would be the same except for the different LO frequencies.



Here is the foil pattern for the component side of the IF66.



Here is the foil pattern for the solder side of the IF66.

Testing the IF44. The IF44 will use the same test setup as the IF66 with the exception of the input signal. Some source of 44-MHz IF signal will be needed. That type of signal can be obtained from a surplus TV tuner taken from a junked TV set. A TV-signal generator can also be used if one is available. Since the IF44 has no local oscillator, no frequency counter or FM radio is needed.

With those differences in mind, the testing procedure for the IF66 can be used with the IF44 except for tests having to do with the IF66's local oscillator; those components are not a part of the IF44.

Note that when the test clip is

removed from R21, the picture should not become garbled; L4 should be adjusted for a 3.5-volt reading on TP2. When adjusting L1 and L2, the target frequencies are 40-47 MHz.

Testing the IF70. The IF70 needs very little setup and should work well enough to see a video image and hear some audio even with no alignment. A digital voltmeter and a source of 70-MHz signal are needed. If no FM video is available, a simple carrier-wave signal will do. Conventional AM video (such as from VCRs) will not work. By using an unmodulated carrier, the settings

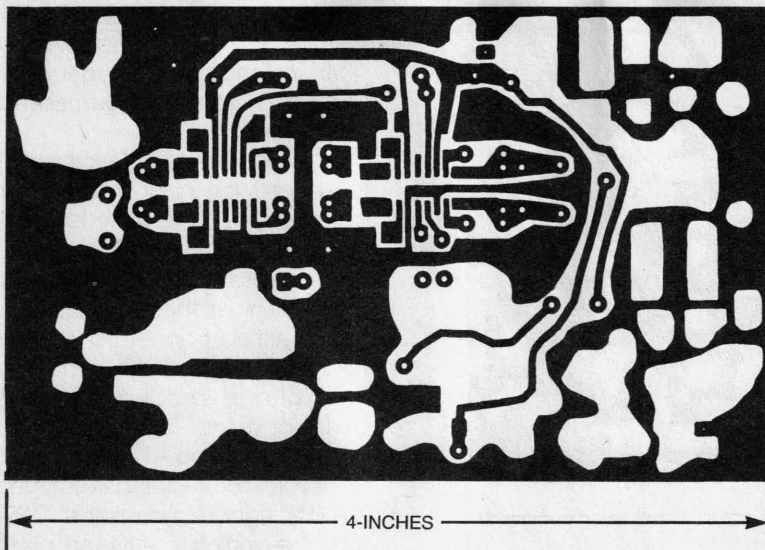
will be close enough so that when an actual FM-video signal is available, the most that will need to be done is a slight readjustment of L1 and C29.

Apply a source of 12-volt power to the board. The chart in Table 3 indicates where to take voltage measurements, as well as the expected reading and its tolerance.

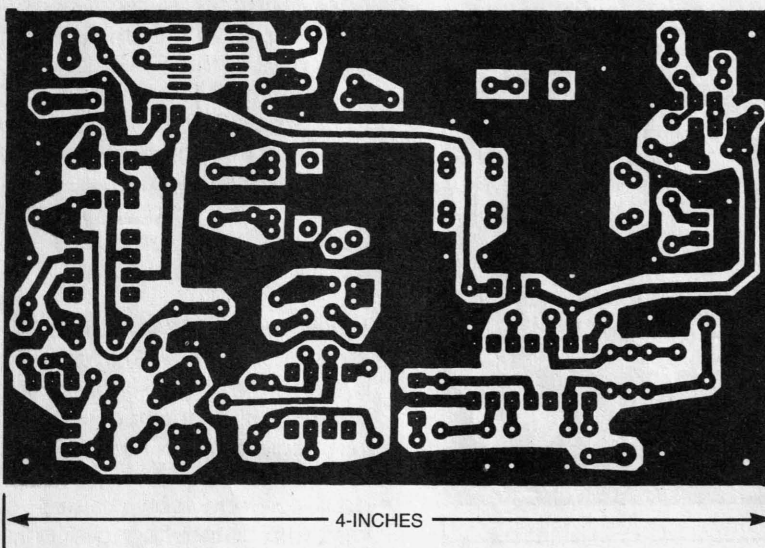
Once all of those voltages have been verified, connect a 2 to 10 mV 70-MHz signal to J1. While monitoring the voltage produced between pins 4 and 5 of IC2 (will usually be between +1 and -1 volts), adjust L1 for a zero-volt reading. Remove the 70-MHz signal and adjust C29 so that the voltage between pin 1 of IC5 and ground is halfway between the maximum and minimum voltages that can be produced while adjusting C29. With that, the initial alignment of the IF70 is done.

When the IF70 is installed in a system, adjust C29 for the best audio (as heard through speakers) and L1 for the best video (as seen on a monitor). The initial alignment will be found to be pretty close to optimum in many cases.

Interfacing the IF44 and IF66. The use of AGC on the RF amplifier stage is strongly advised when using the Video IF with a downconverter. While input signals up to 10 mV or so can be handled, a 30-dB gain on the downconverter can cause IF overload with an input signal of only 300 microvolts. For best results, it is recommended that the AGC output from TP6 be connected to the gain-control input of the downconverter. A capacitor between 10 and 100 μ F should also be connected between the AGC signal and ground. That will provide filtering and a time constant. Failure to do that will result in picture breakup on strong signals; horizontal "pie crusting"; horizontal lines; and an otherwise very noisy, poor-quality picture. It is best to experiment with the capacitor value for best results in a particular situation. The capacitor should ideally be non-polarized, as AGC voltages can vary from +1 volt at maximum gain down to -3 volts. In practice, most good quality aluminum electrolytic capacitors of the polarized type will function up to somewhat more than



Here is the foil pattern for the component side of the IF70.



Here is the foil pattern for the solder side of the IF70.

1 volt of reverse voltage without damage. In that case, make sure that the POSITIVE lead is connected to ground. If you are uncomfortable with that approach, use a non-polarized unit, a Mylar device, or create a non-polarized device by wiring two polarized units back to back in series. Downconverters of various manufacture might need additional modifications that depend on circuitry. That is beyond the scope of this article; it is left to the experimenter.

You should adjust R18 so that the AGC voltage starts to fall from +1 volt toward the negative levels when a signal of about 5 to 10 mV is present at the IF input. That will let

the downconverter run at full gain at up to 100-300 microvolts.

The Video IF has been designed to interface with downconverters that have appeared in past issues of **Electronics Now**, including the May-June 1996 and September 1992 issues.

With the various versions of the Video IF, you can work with just about any type of video signal available today. As you experiment with them, new uses will suggest themselves as you become familiar with their capabilities. Ω



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PLL Techniques

— a practical guide

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Recently, I had occasion to work on several different phase locked loop systems. One system was relatively straightforward, but two of the systems had to utilize a heterodyning process, and these proved to be anything but straightforward. More recently, I came across a statement made by a Mr. R. E. Funk of RCA (*COS/MOS Digital Integrated Circuits, Solid State Handbook SSD-203B, page 460*). Mr. Funk says: "The disadvantage of the heterodyne technique is that, because a second reference crystal and a mixer are needed, the designer may be afforded an unsought opportunity to demonstrate his process in eliminating spurious beat and sum frequencies." This is the understatement of the year! But Mr. Funk certainly knew what he was talking about.

The purpose of this article is to share some of my experiences. There is a definite shortage of practical phase

locked loop material. The publication just mentioned and the *Phase Locked Loop Systems Data Book* put out by Motorola contain practically all the useful data I could find. I recommend that the serious experimenter obtain these publications.

There have been various ham articles on phase locked loop systems, but these have been somewhat limited in content and spread over a number of publications. I propose, therefore, for the purpose of completeness, to begin at the beginning and build up to a practical system. Hopefully, this discussion on the various pitfalls will save you a few headaches and frustrations!

First of all, you may ask, "Why do I want a phase locked loop system at all? What is the advantage of it?" To answer this, it is necessary to discuss the shortcomings of some other systems. For example, an FM transmitter may generate the FM at a low frequency, 3 MHz being typical, and double and triple many times until the desired output frequency is obtained. This is fine, except for one

thing — the output of the transmitter may contain many spurious signals such as 6 MHz, 9 MHz, 12 MHz, and so on. This is the reason the 2 meter transmitter contains so many tuned circuits — regulations and good practice say that these products must be kept at home.

In many cases, it is necessary to obtain a local oscillator injection voltage that is too high in frequency for a stable vfo. The usual method is that a low-frequency vfo is mixed with a higher-frequency crystal oscillator output, and the sum or the difference product is utilized. But this method also provides a lot of spurious outputs, and, unless the mixer output is adequately filtered, it will result in the reception of all sorts of unwanted signals. As an example, a 5 MHz vfo may be mixed with a 33 MHz crystal in order to obtain a 38 MHz local oscillator signal which, when used with a 12 MHz i-f, will allow 50 MHz operation. In this instance, the sum product has been used, but there is also a difference product. 33 MHz less the vfo frequency of 5 MHz equals

28 MHz. Unless this signal is properly filtered, you might find 40 MHz commercial stations coming through in addition to the 6 meter signals. To make matters worse, harmonics of the vfo will also be present in the local oscillator chain output, one of which will be 40 MHz, only 2 MHz away from your wanted output.

Now, in the case of the FM transmitter mentioned earlier and in the local oscillator system just described, an oscillator operating right at the output frequency would eliminate the need for multiple tuned circuits and elaborate filters, because a properly designed oscillator will be free of all spurious signals except harmonics, and, usually, these are sufficiently removed in frequency not to be troublesome. Unfortunately, unless it is possible to lock the high-frequency oscillator to a stable source, you will have traded a spurious signal problem for a frequency stability problem. This is where the phase locked loop enters the picture. The system allows the high-frequency oscillator to be locked to a stable source. If crystal control is required, a 154 MHz oscillator, for example, may be locked to a 1 MHz or a 10 MHz crystal oscillator. However, to supply many different local oscillator frequencies, many crystals must be used or, alternatively, a synthesizer inserted into the circuit (more on this later).

The local oscillator or the FM transmitter may also be locked to a lower frequency vfo. But, before I discuss this, refer to Fig. 1, and see how a vfo may be locked to a crystal. Incidentally, when a vfo is locked in this manner, it is called a vco — voltage controlled oscillator. Referring to Fig. 1, the vco feeds output into a kind of mixer which is usually termed a phase detector (0 detector). Actually, many phase detectors are simple mixers. At the same time, the crystal oscil-

lator, which is now called the reference oscillator or sometimes the clock, also feeds output to the mixer or phase detector. When there is a difference in frequency between the two input signals, certain mixers will deliver a dc output which is either negative or positive in polarity, but will have no output if the two input frequencies (and phase) are the same. It is this + or - output voltage which distinguishes a phase detector from a simple mixer. The dc output from the phase detector may now be used to change the capacitance of a variable-capacitance diode which is part of the vco tuned circuit and thus vary the vco output frequency. If the circuit is properly designed, the dc voltage (usually known as the control or error voltage) will move the vco frequency toward the reference frequency. When the two frequencies are the same, there will be no error voltage, and the vco frequency will stabilize. If the vco moves higher in frequency, the phase detector will deliver a + error voltage, which will increase the diode capacitance and lower the frequency. Conversely, if the vco drifts lower in frequency, a negative error voltage will move the vco higher. Thus the vco is locked to the crystal.

Of course, there is little advantage in using the simple circuit just described. It would be simpler to use the crystal oscillator directly. However, the circuit does illustrate the principle. What if the reference oscillator was "dirty," that is, it contained a number of unwanted spurious products? Well, just what we would like to happen would: The vco would clean up the signal. Because the vco is in itself a sine wave oscillator, kept on frequency only by occasional squirts of dc, it should have pure output. In practice, this is not always the case, but a properly designed system will certainly reduce spurious signals, usually to an acceptable level.

Incidentally, the vco may also be a multivibrator type, delivering square wave output. The type I propose to discuss, because it may be used with SSB receiving and transmitting equipment and thus must contain little phase jitter, will be the sine wave generator.

The block diagram shown in Fig. 2 illustrates a scheme utilized in many modern CB sets where, because of the large number of crystals required to give complete channel coverage, a system that mixes a smaller number of crystals is preferred. The outputs of oscillators #1 and #2 are fed into a conventional mixer. Either the sum or the difference product is selected, whichever is convenient, and the resultant product is fed to the phase detector. Of course, a double-balanced-type mixer has to be used here to suppress the two original frequencies, otherwise the vco will not know which signal to lock to. However, a "dirty" signal has now been made clean.

Although Fig. 2 is a practical system used by many manufacturers, it still requires something like 11 crystals in the case of a 23-channel CB set. Generally speaking, crystals are a nightmare of supply problems to most manufacturers of original equipment. Fortunately, there is another way. Referring again to Fig. 1, let's say that in some mysterious manner we are able to "rubber" the crystal 1 MHz higher in frequency. What will happen to the vco? It will follow right along and stay locked to the reference. So the problem really is how to "rubber" the crystal to where we would like it.

Referring now to Fig. 3, it will be seen that a new building block has been added between the vco and the phase detector which is labeled divide-by-N. For the moment, let's have $N = 2$. A simple flip-flop, costing less than 50 cents, will fill this

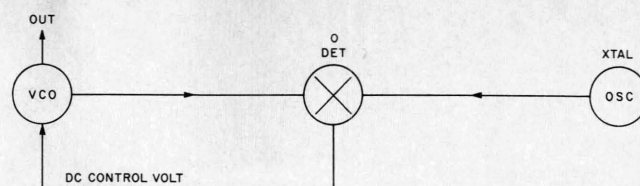


Fig. 1. This block diagram shows how a vco may be locked to a crystal oscillator. Both the vco (herein called a vco) and the crystal oscillator feed signal to the phase detector. The phase detector generates a dc voltage which "steers" the vco on to frequency whenever there is a difference in frequency between the two oscillators.

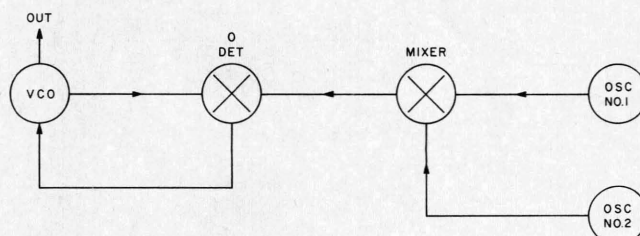


Fig. 2. When a high-frequency local oscillator is required, a vco (osc. 1) may be mixed with a crystal oscillator (osc. 2) to obtain the necessary high-frequency output. This signal will contain many unwanted spurious outputs. However, if the "dirty" signal is only used to lock the vco, the unwanted products will be filtered out.

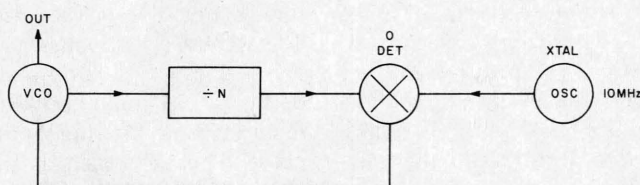


Fig. 3. In this system, unlike that shown in Fig. 1, the vco may operate on a different frequency from the crystal. For example, if the $\div N = 2$, the vco may operate at 20 MHz. See the text for more on this.

block. Because the output of the vco is connected to the flip-flop input, the vco output is divided by two, and the result is given to the phase detector. The block diagram shows that in this instance the reference frequency is 10 MHz. Had the vco been generating a signal around 10 MHz, the vco would divide this by two and feed a 5 MHz signal to the phase detector. Because the 10 MHz reference signal is considerably different in frequency from the 5 MHz received at the input port, a large error voltage will be produced, and the vco will be retuned until both ports receive 10 MHz signals. This means that the vco will now be delivering output at 20 MHz. If we add

a second flip-flop in tandem with the first so that $N = 4$, the vco will move to 40 MHz in order to reach lock. Thus, merely by changing the value of N , a number of vco frequencies may be obtained which are all locked to the reference frequency. The manufacturers of integrated circuits have come to our aid with a number of inexpensive flip-flops in one package. The package may readily be programmed to give a number of division ratios. For example, the RCA CD4018 will divide by any number between 2 and 10. When a number of counters are combined in this manner, the simple flip-flop counter becomes a synthesizer.

Referring again to Fig. 3,

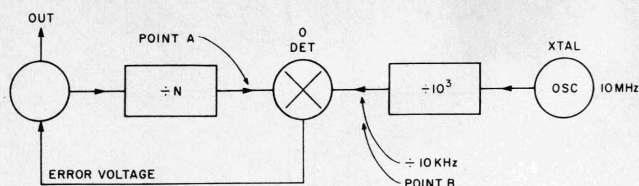


Fig. 4. The system shown in Fig. 3 allowed the vco to move only in 10 MHz steps. By dividing the reference oscillator down to a lower figure (for example, 10 kHz, point B), then the vco may supply outputs only 10 kHz apart.

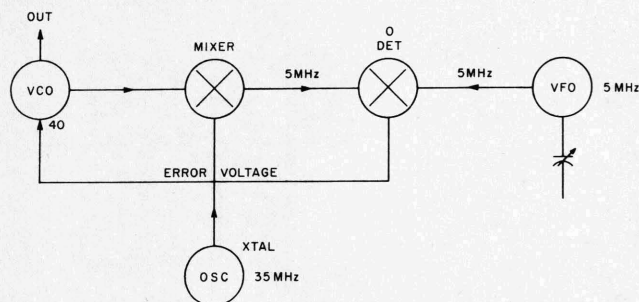


Fig. 5. In this scheme, the vco is heterodyned against a crystal oscillator to produce a difference signal in the same frequency range as the vfo. When both the heterodyned output and the vfo output are fed to the phase detector, the vco will lock to the vfo and have the same stability.

it will be seen that the vco output can only move in 10 MHz steps, because you are only able to divide by whole numbers. However, it will be readily seen that, if the reference frequency is lowered to 1 MHz, the vco will move in 1 MHz steps. Thus, if a 10 kHz channel spacing is required, the reference frequency must be 10 kHz. A 10 kHz crystal is kind of expensive, but, fortunately, you can retain your 10 MHz or 1 MHz crystal and simply divide it down by using a number of inexpensive flip-flops. A few type-7490 ICs will do this job for less than a dollar. This scheme is shown in Fig. 4. Input to the phase detector at point A in Fig. 4 will not always be 10 kHz, but it will always be 10 kHz when the system is in lock. When there is an 11 MHz signal at point A, an error voltage will immediately develop and, regardless of the value of N, will move the vco frequency until 10 kHz appears at point A.

In the above explanations, it was said that only a dc error voltage was obtained from the phase detector, but

this is not always the case. The RCA CD4046 phase detector does give a fairly clean dc signal, but the output still requires some filtering. Some phase detectors, such as the Motorola MC4044, deliver a series of pulses at the output terminal, and the output may require considerable filtering. Each type of phase detector has its particular usages. Currently, I am using both types in newly-designed equipment. The divide-by-N part of the circuit is beyond the scope of this article.

The systems outlined above have one serious disadvantage. While they operate well if fixed-channel operation is required, such as is used in the 2 meter band, they do not lend themselves to the sweeping or the searching across the spectrum. To properly tune SSB, one must be able to tune to within 10 Hz, not 10 kHz. To find a buddy due to come up somewhere between 14230 and 14250 requires a vfo knob, not a number of synthesizer knobs. At first thought, it would seem that the simple answer is to replace the reference oscillator with a vfo. I

wish it was that easy. Unfortunately, this scheme suffers from the same disadvantage you have when a vco is multiplied to obtain a higher local oscillator frequency; the frequency instability is multiplied by the same factor.

In our case, if the divide-by-N = 8, the frequency instability of the vfo will also be multiplied by eight. What was a 100 Hz drift per hour now becomes an 800 Hz drift per hour! Additionally, each time the divide rate is altered, not only does the frequency drift rate change accordingly, so does the amount of frequency covered by the dial. This makes for difficult dial calibration unless a digital readout is used.

The Heterodyning Method

Just as you can heterodyne a vfo to a new frequency by beating it against a crystal oscillator, so you use this technique in the phase locked loop system. The basic scheme is shown in Fig. 5. In this example, a 40 MHz local oscillator is required for a 6 meter receiver. Output from the 40 MHz vco is fed into an ordinary mixer together with signal from a 35 MHz crystal oscillator. The 5 MHz difference output from the mixer is then fed to one input port of the phase detector. Output from a 5 MHz vfo is simultaneously fed into the other input port of the phase detector. If a difference between the two input signals exists, an error voltage is dispatched to the vco and its frequency moved until lock is obtained. When the vfo is moved, this, too, generates an error voltage, and the vco is forced to follow. If the vfo is moved 10 kHz, the vco will move 10 kHz. If the system is properly designed, the vco output will be clean and contain none of the 35 MHz component, vfo harmonics, or the like. And so it appears that the heterodyning system is the panacea to our problems, and, in some cases, this is so. However, in many other instances, the problems are

just beginning.

In the example shown in Fig. 5, when it's first turned on, before lock is obtained, the vfo output frequency may be anything within the range of the system. If the phase detector output is able to vary from 0.5 volts to 4.5 volts, which is typical (with 2 volts being the nominal center frequency), the variable capacitance diode may have its capacitance changed by as much as 100 pF. This change may readily sweep the vco as much as 20 MHz before lock is obtained. If, referring to Fig. 5, perchance the vco is moved down to 30 MHz, this frequency when mixed with the 35 MHz oscillator will also produce a 5 MHz difference signal at the mixer output. Naturally, the loop will try to lock to this signal. However, as the error voltage will now be of the wrong polarity, lock will not be obtained, the phase detector will be thoroughly confused, and the vco will settle at some highly unstable frequency below 35 MHz. Fortunately, in the example given, the image is sufficiently removed in frequency from the wanted signal. The vco swing may be sufficiently restricted so that it is not capable of running into this highly dangerous area. But, in instances where the difference frequency is small, this phenomenon becomes a real and serious problem. There are various ways one may overcome this problem. One method makes use of the fact that, when the vco is tuned to the image frequency, the error voltage will become highly negative with respect to the nominal frequency (in this instance only; it may become positive in other instances). When the high negative voltage is produced, it may be caused to trigger a ramp generator which, in turn, forces the lock line in the other direction. Such circuits, in my experience, also contain numerous gremlins and tricky elves.

Another method may

make use of a tuned circuit lightly coupled to the vco tuned circuit. If the accessory tuned circuit is tuned to the dangerous frequency area, it will receive a signal when the vco is in that general area. The signal may be rectified and used to ramp the lock line.

Some phase detectors will lock to a harmonic of the wanted signal voltage. One of the phase detectors in the RCA CD4046 package (there are two) displays this annoying characteristic. It's quite disconcerting to obtain good lock, but on the wrong frequency.

Whatever the system employed, the heterodyne method almost always requires that the mixer output be filtered to eliminate the vco and crystal oscillator outputs and also the unwanted image frequency. If the difference frequency is used, a simple low-pass filter may be sufficient, but, often as not, a quite elaborate filter must be installed between the mixer and the phase detector.

And this brings us to a new problem, also quite common and often serious. If the vco is able to swing sufficiently far in one direction so that the mixer output frequency is greater than the low-pass filter cut-off frequency, the filter will prevent the signal from reaching the phase detector. The phase detector, in alarm, will send a panic voltage to the variable-capacitance diode, which, in turn, is sent even further away in frequency. The system has latched up and will stay latched until a ramp is activated or the supply voltage is removed and replaced.

From all of the above, it would seem that the easiest thing to do is to limit the vco swing so that the dangerous latch-up area cannot be entered. Indeed, this is mandatory in most cases. The vco is often preset. That is, each time the MHz bandswitch on a receiver is moved, for example, an additional diode across the vco tuned circuit

has the voltage across it altered so that the vco is "steered" near the wanted frequency.

If the vco swing is too restricted, it will be found that lock will not be fully obtained near the band edges. Instead, the vco will "skip" cycles and never quite make it.

Other problems, not only peculiar to the heterodyning method, but also to the other methods outlined at the beginning of this article, are caused when the vco is allowed to swing beyond the "capture range" of the system. With each phase detector and PLL system there are extremes beyond which lock cannot be obtained; however, if they are forced into lock, lock will be maintained.

The capture range, to a major extent, is also dependent upon the loop bandwidth, which is tied up with the capture time. This is highly dependent upon the type of filtering required in the lock line. Because the error voltage to the variable-capacitance diode may contain spurious voltages, especially the reference voltage, it must be filtered. If the reference frequency is in the audio range, for example, filtering may become quite troublesome. If the filtering is inadequate, the vco signal may be modulated by the spurious components. Often, if the filtering is good, the lock-up time and thus the capture range may suffer. Filters must often be skillfully designed, and this design may well be one of the more difficult parts of the system. Capture range is exactly what it says — the frequency range over which an out-of-lock system will lock. It will be apparent that, once lock is obtained, moving the frequency of either of the oscillators at a rate slower than the time required to obtain lock will cause only a small error voltage, and correction and lock will be easily maintained. If, however, lock-up time is very slow, the error voltage change may arrive too

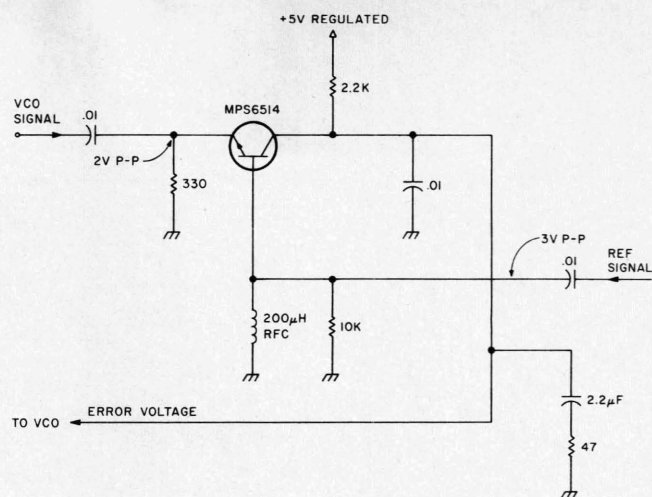


Fig. 6. A simple phase detector. This detector has poor lock capture ability. Various other phase detectors are described in the article.

late, and the system may drop out of lock. This is particularly true when a different crystal is switched into circuit, resulting in a large frequency shift and possibly a temporary loss of signal at the input to the phase detector. Before the error voltage can arrive, the system may have dropped out of lock and even latched up.

Contrary to general belief, the signal to the phase detector must be fairly "clean." Some phase detectors do practically ignore spurious signals mixed with the wanted signal, but others are driven to a cantankerous behavior. The previously-mentioned CD4046 phase detector contains two different types of detectors in the one package, only one of which is affected by "dirty" signal frequencies (phase detector #2). On the other hand, phase detector #1 readily accepts signals amid static, shot noise, and other garbage, but it readily responds to harmonics. In instances where the signal is "dirty" and the system could lock to harmonics, the MC4044 has proven an excellent choice.

The #2 detector in the CD4046 package, because its output is not a series of pulses, is very readily filtered, usually only requiring a resistor and a capacitor unless the reference frequency is low. Even then, the filter network

will be simple and inexpensive. Phase detector #1 and the MC4044, because they deliver a series of positive- or negative-going pulses, require much more elaborate filtering and may even require active two-step filter systems.

From reading this, it may seem that only two phase detectors are available on the market, but this is far from the case. Fig. 6 shows a simple phase detector. This type of detector is suitable for many applications, but it has a very small capture range unless ramped. The sample-and-hold-type detector has very clean output, usually utilizing two or more field effect transistors. The diode-type double-balanced mixer or modulator makes an excellent phase detector, but it has very limited capture range. The digital types, such as those mentioned earlier, because they have been especially designed for phase detector application, do make for simple detection. Unfortunately, these devices have a limited frequency of operation, especially the CD4046, which is quoted at 1.2 MHz upper frequency. The MC4404 is quoted at 8 MHz. This means that, in many circuits, both the signal and the reference frequencies must be divided lower in frequency merely to get within the operating range of the device, an unfortunate com-

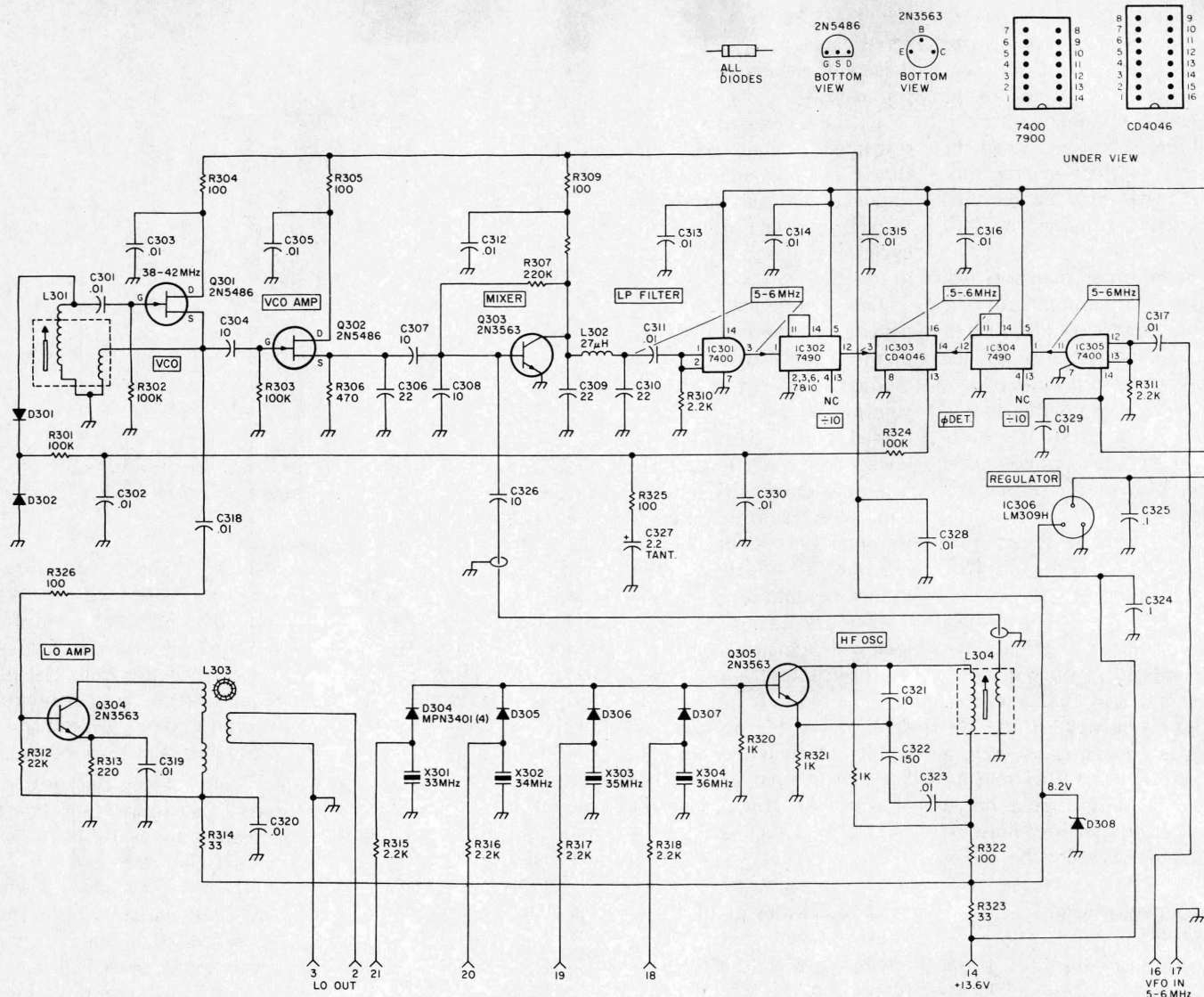


Fig. 7. The schematic diagram allows a vco covering the frequency range of 38 MHz to 42 MHz to be locked to a vfo covering the frequency range 5 MHz to 6 MHz.

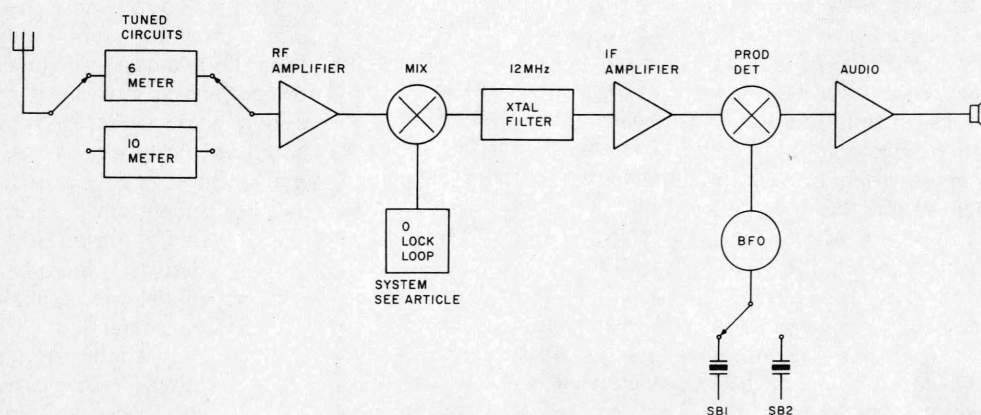


Fig. 8. The phase locked loop system shown in Fig. 7 and covered in the article may be used with a receiver which will allow operation on both the 10 and the 6 meter bands. Although the system uses a high-frequency local oscillator, the stability is equal to and determined by the vfo which operates over the frequency range 5 MHz to 6 MHz. The block diagram shows that it is only necessary to change the front-end tuned circuits to change bands.

plication.

Please note that only some of the more serious problems

have been touched upon in this article. There are other problems, such as phase jitter,

that space will not allow to be discussed. The more serious designer will be con-

fronted with a host of others. This is especially true if he possesses a spectrum analyzer and can more readily see what is happening. In many ways, life before the invention of the spectrum analyzer was less hectic!

A Practical Phase Locked Loop System

The PLL system shown in schematic form (Fig. 7) is derived from the block diagram of Fig. 5. In this particular system, a vco operates over the 38 MHz to 42 MHz frequency range, allowing either 10 meter or 6 meter operation with an i-f frequency of 12 MHz. The vco, Q301, is an FET Hartley-type oscillator, the output of which is fed to an amplifier

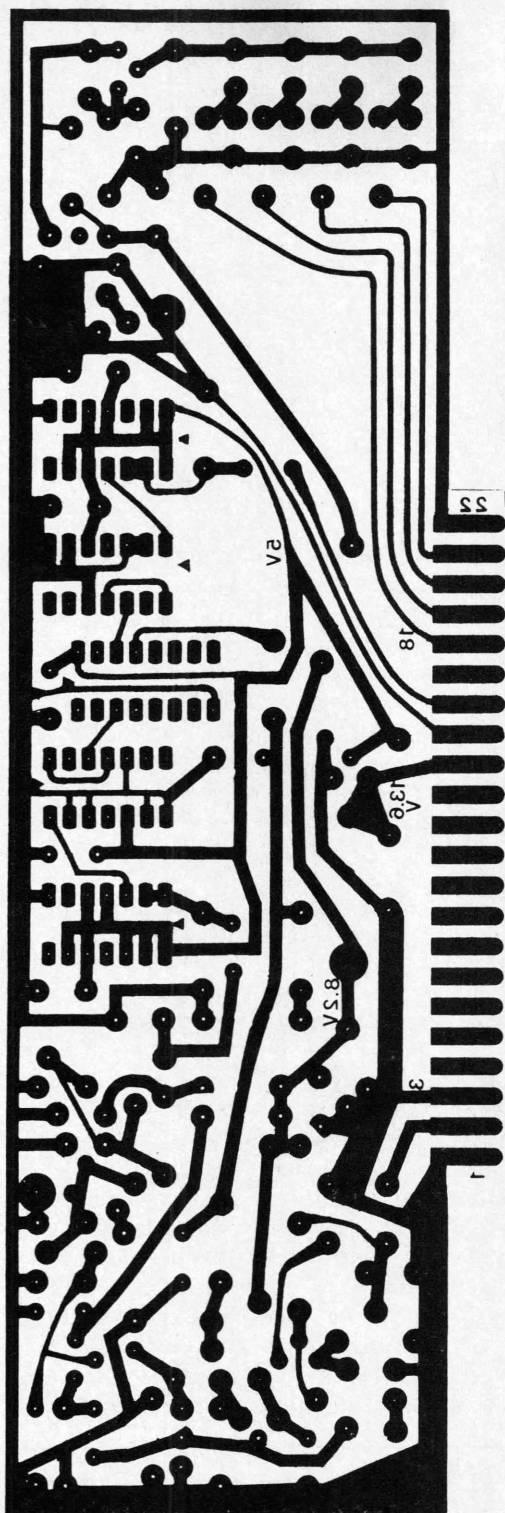


Fig. 9. PC board.

stage, Q304, and thence to the receiver mixer and also to vco amplifier Q302. The output from Q302 is connected to the PLL mixer, Q303. A high-frequency crystal oscillator, Q305, is operated together with one of four diode switched crystals. From 8 to 12 volts applied to one of the diodes at one time will turn

on the appropriate diode. Output from the high-frequency oscillator via IFT L304 is fed to the PLL mixer, Q303.

All outputs from the mixer, except the difference between the two oscillator signals, are suppressed by C309, L302, and C310. The wanted frequency, covering the range

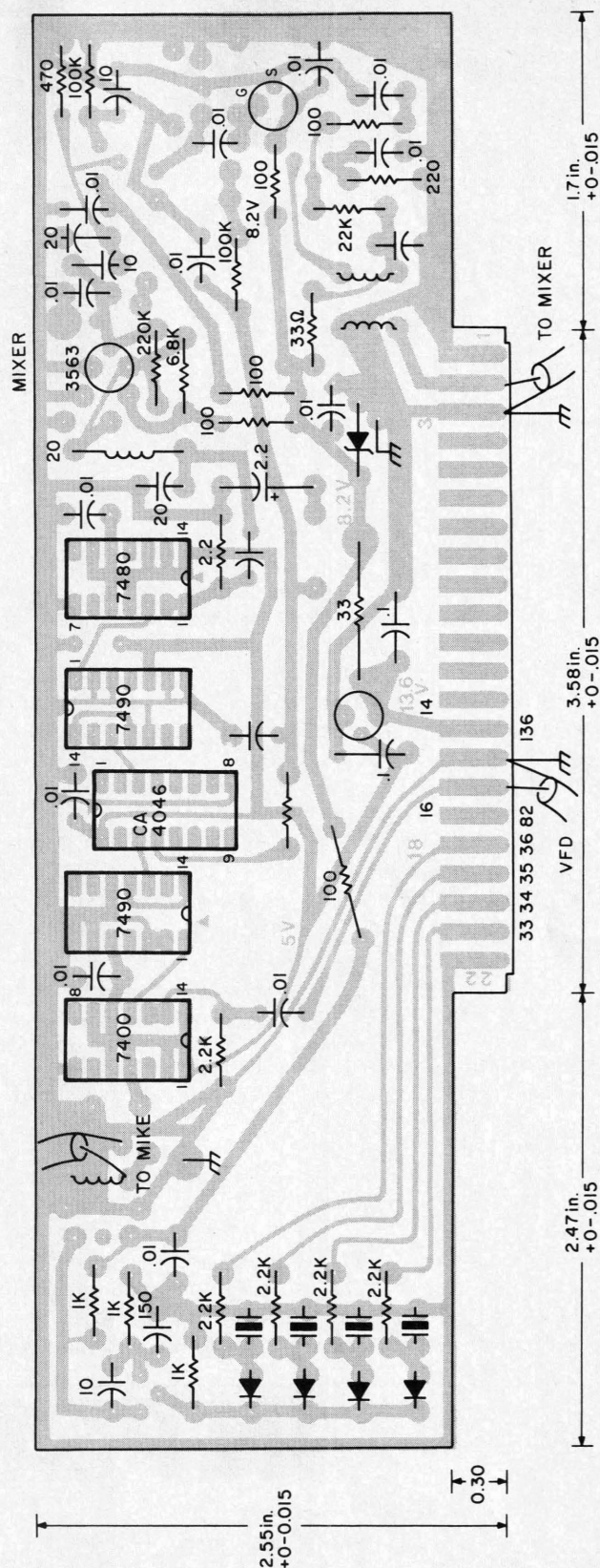


Fig. 10. Parts layout.

5 MHz to 6 MHz, is shaped and amplified by one section of a 7400 quad gate before being applied to the divide-by-10 counter. The counter is necessary because the 5 MHz to 6 MHz signal is too high

for the phase detector, IC303. Thus the phase detector receives a signal in the 0.5 MHz to 0.6 MHz range.

A vfo covering the range 5 MHz to 6 MHz, delivering a

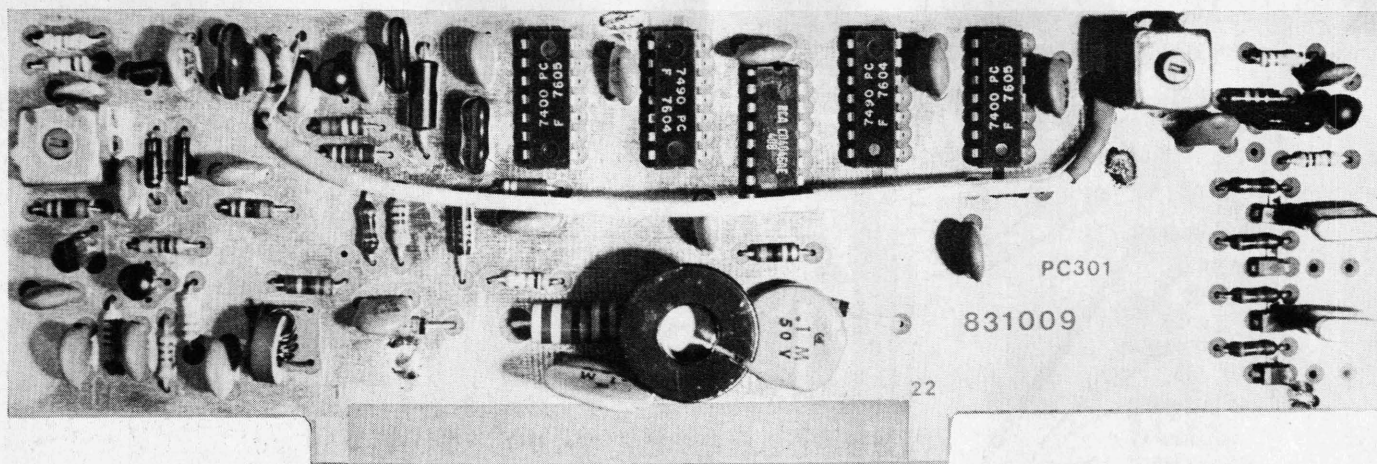


Photo A.

sine or square wave with an amplitude of at least 0.3 volts rms, is also shaped and amplified by a section of a 7400 quad gate, IC305. The output from this device is also divided by 10, and the resultant 0.5 MHz to 0.6 MHz signal is applied as the reference to the phase detector. The error voltage output from the phase detector, pin 13, is filtered by components R324, C330, R325, C327, and C302 before being applied to the variable capacitance diodes, D301 and D302. The diodes shown are somewhat expensive KV1501s, but it has been found that rectifier diodes (1N4001s) functioned quite well in this circuit. It may be necessary to select diodes, and they may be somewhat noisier (a white noise effect).

The vco coil was wound on a Micrometals form #L43-6-CT-F-5. The primary consists of 6 turns of #30

enameled wire spaced the diameter of the wire. The secondary consists of two turns wound between the turns at the bottom of the primary. It is necessary to observe the correct polarity of the windings if oscillation is to be obtained. If the Micrometals form is not available, other forms may be used. Temporarily disconnect the lock line from R301, feed 2 volts into the resistor, and adjust the coil slug until oscillation is obtained at about 40 MHz. This will get the coil into the correct range. IC306 is a three-terminal regulator which drops the supply voltage to 5 volts required by the ICs.

Transformer L304 is wound on the same form as L301. The primary consists of 12 turns of #30 closewound and the secondary of 4 turns closewound over the bottom of the primary.

To adjust the loop, first

adjust L304 until reliable oscillation is obtained on all four crystals. (Note that, since Photo A was taken, the position of the crystals and diodes was reversed to obtain better operation. The schematic is correct.) While observing the output of L303 on a high-frequency oscilloscope and while measuring the frequency, adjust the slug in L301 until the voltage on the lock line is about 2 volts when the vfo is set at 5.5 MHz.

The frequency should remain stable and not move unless the vfo is moved. To check for proper operation, short the lock line to ground for a moment. The vco should immediately return to frequency.

The wideband transformer L303 is wound on an Indiana General Q1-type core and consists of 4 turns trifilar wound using #26 enameled wire. Almost any core suit-

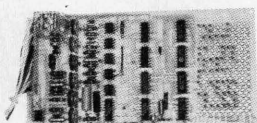
able for wideband work may be used here.

The oscillator and PLL system, if used with a 12 MHz i-f, will have a "birdie" in the receiver when the vfo is at the 6 MHz part of the range. This is due to the second harmonic of the vfo falling in the i-f range. This is readily removed if the crystal and vfo are moved a few kHz to get the harmonic out of the passband.

The practical PLL system described, while not of use to everyone, has been constructed and operated in a receiver covering 6 and 10 meters. It may serve to assist others building similar equipment. Other similar systems have been designed covering a wider frequency range. In these systems, the mixer was replaced with a double-balanced type to remove the two oscillator signals. Other designers will have their own ideas on this. ■

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Low-Pass Filter Primer

—tells all

Although my interest in active low-pass filters is due mainly to my interest in RTTY, they can be found in other places as well—transmitters, speech amplifiers, synthesizers, etc. (If you are interested in active bandpass filters, see my article in the September, 1977, issue of 73 Magazine. That article also provides some background information that is of use in connection with low-pass filters.)

An active filter is a filter which uses an amplifier, usually an operational amplifier integrated circuit such as a 741, in an RC circuit which achieves the same performance as would otherwise require an LC circuit. As opposed to the LC filter, an active filter is usually cheaper to build since it does not require expensive and bulky inductors, and is often easier to design and trim as well. In

case your knowledge of filters needs a little brushing up, the following discussion will bring you up to date and introduce some of the basic concepts of filters.

Single-Stage Filters

A passive low-pass filter—that is, one not using an amplifier—can be built with either just RC components or with an LC circuit. Fig. 1 shows the diagrams of several such simple low-pass filters, while Fig. 2 shows their frequency response. The simplest is the single-stage RC filter, Fig. 1(a). Because it has just one reactive component, a capacitor, it is called a one-pole filter. Its frequency response curve is the top curve in Fig. 2.

Looking at the top curve in Fig. 2, we see that the far left of it approximates a straight line, and the far right of it also is like a straight line. If we take a ruler and draw two dashed lines as shown, continuing the straight portions of the curve toward the middle, they meet at what is called the *cutoff frequency* or *corner frequency*—this is the frequency where the two straight lines meet at a corner. To keep things simple, this frequency response drawing assumes a corner frequency of 1 Hz, but the

curve would have the same shape regardless of what the frequency is. For instance, a 50 Hz filter would look the same, but all frequency readings shown along the bottom would be multiplied by 50.

The plain RC filter's response is down 3 dB at the corner frequency; at half this frequency it is down 1 dB, while at twice that it is down 7 dB. Once it becomes straight, beyond 4 Hz or so, it continues to drop 6 dB for every doubling of frequency (this is called *6 dB per octave*), which is the same as 20 dB for every *decade* (a decade is a 10 to 1 ratio of frequency).

When two stages of RC filtering are put together, as in Fig. 1(b), they interact to some extent and produce strange results. But if it is done right, or if the two stages are separated by an amplifier so the second does not load down the

output of the first, then the result is the bottom curve of Fig. 2. Here the corner frequency is the same, but the response at this frequency is down 6 dB, 3 dB for each stage. It is down 2 dB at half the frequency, 14 dB at twice the frequency, and it drops at the rate of 12 dB per octave or 40 dB per decade, exactly twice the rate for the single-filter stage. This filter has two reactive components, both capacitors, and so it is called a two-pole filter.

Another two-pole filter is the LC filter of Fig. 1(c) which has two reactances, one inductor and one capacitor. If designed just right, it, too, has a response curve like the bottom one in Fig. 2. But depending on the Q of the filter—the amount of resistance in the circuit introduced by the LC components—the LC filter can have other response curves as well, as shown in Fig. 3. A lossy cir-

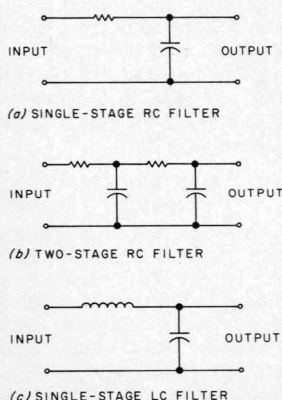


Fig. 1. Some simple passive filters.

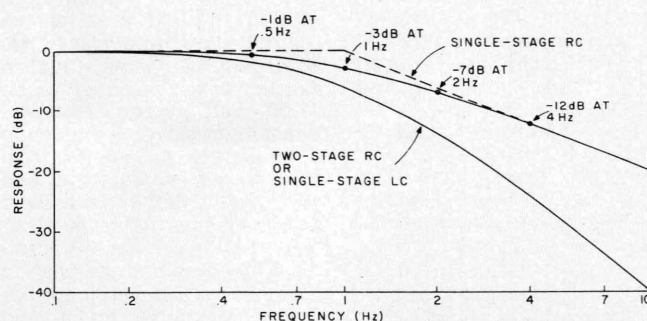


Fig. 2. Passive filter frequency response.

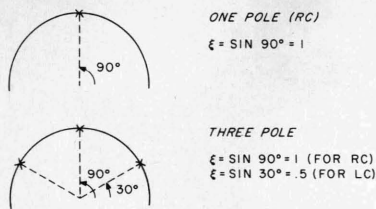


Fig. 6. Pole locations for odd numbers of poles.

see that for the four-pole filter, the poles are $22\frac{1}{2}$ and $67\frac{1}{2}$ degrees above the horizontal, so the two values of ξ needed are $\sin 22\frac{1}{2}^\circ = 0.383$, and $\sin 67\frac{1}{2}^\circ = 0.924$. Keeping in mind how the poles have to be located on the circle, you could draw your own pole locations for as many poles as you want. For every two poles you would then need one LC circuit.

Although Fig. 5 shows only even numbers of poles, it is possible to build Butterworth filters having odd numbers of poles. For instance, a single RC stage has a single pole, at the very top of the semicircle as shown in Fig. 6, at the top. In this case, the angle above the horizontal is 90° . A three-pole filter, shown at the bottom of Fig. 6, would have three poles still evenly spaced, 60° apart. The top pole would be produced by an RC filter stage, while the other two poles would be produced by an LC filter having $\xi = 0.5$.

Although these pictures of poles are interesting and provide an easy way of remembering where they go and how to calculate the

required value of ξ for any desired filter, it is easier to consult a table like Table 1 for the exact values needed.

So far we have been discussing only RC and LC filters. But, as mentioned before, LC filters are hard to adjust, expensive, and often large. Fortunately, an active filter, using just RC components plus an operational amplifier integrated circuit, can produce the same response as an LC filter.

Active Filters

Though there are several filter circuits which can be used, that of Fig. 7 is probably the simplest, using just one operational amplifier IC, three resistors, and two capacitors.

The design procedure is to start off by picking a convenient value for capacitor C_2 ; most audio designs use values of perhaps 0.01 or 0.1 μF . Next, choose how much gain (G) you want the filter to have at dc and low frequencies. Best operation will occur when G is less than 10, although it can be made as high as 100 when the filter is designed for

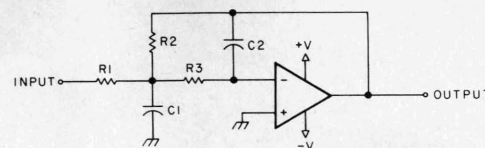


Fig. 7. Diagram of one active low-pass filter stage.

low frequencies and ξ is large.

Once C_2 and G are chosen, find the other component values from the following equations:

$$C_1 = \frac{C_2 (1 + G)}{\xi^2}$$

$$R_2 = \frac{\xi}{2\pi f C_2}$$

$$R_1 = \frac{R_2}{G}$$

$$R_3 = \frac{R_2}{1 + G}$$

Although it would certainly be nice to use the exact calculated values in building your circuit, such accuracy is not usually needed; commercial values are usually good enough since the exact operating frequency or damping of low-pass filters is not usually critical.

Nevertheless, if you find you want to play with the circuit values a bit after they are calculated, here are some hints. Changing R_1 , within a fairly large amount, affects only the gain, since $G = R_2/R_1$. So making R_1 smaller will increase the gain, while making it larger will decrease the gain. But keep the gain below 10 if ξ is small.

Sometimes, it may be handy to change the value of C_1 if it comes out to be different from what is

available. C_1 and R_3 are related and can be changed as long as their product stays the same. For instance, if C_1 is calculated at 0.02 μF and R_3 is 590 Ohms, you could cut C_1 by two to 0.01 μF and double R_3 to 1180 or 1200 Ohms; the circuit performance would stay the same, since dividing one by 2 and multiplying the other by 2 keeps their product the same.

When building active filters, you must use good components. Do not use disk capacitors, even for testing; polystyrene or polycarbonate capacitors are best, though plain tubulars or mylar capacitors will work well, too. Also use good op amps, such as the 741, 1458, 5558, or 747; I have had bad luck with the LM3900 in some filters. If you need two or more op amps for the same filter, use separate ICs; I have found some undesirable interactions between two amplifiers in the same IC when used in high-gain filters.

A Practical Example

For use as a low-pass filter for RTTY, I needed a low-pass filter with a cutoff frequency of 50 Hz. I decided I wanted a four-pole filter which would cut off rather quickly, so I went for two active filter stages, as follows:

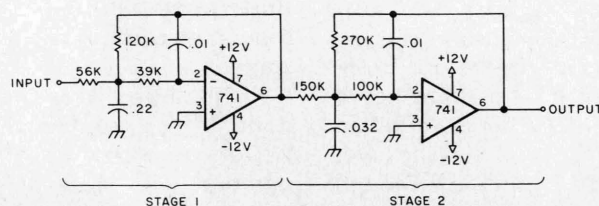


Fig. 8. Complete four-pole low-pass filter.

No. of poles	Required values of ξ			
1	1.0 (RC)			
2	0.707			
3	0.5	1.0 (RC)		
4	0.383	0.924		
5	0.309	0.809	1.0 (RC)	
6	0.259	0.707	0.966	
7	0.223	0.623	0.901	1.0 (RC)
8	0.195	0.556	0.831	0.981
9	0.174	0.5	0.766	0.940 1.0 (RC)
10	0.156	0.454	0.707	0.891 0.988
11	0.142	0.415	0.655	0.841 0.959 1.0 (RC)
12	0.131	0.383	0.609	0.793 0.924 0.991

Table 1. Values of ξ for Butterworth low-pass filters.

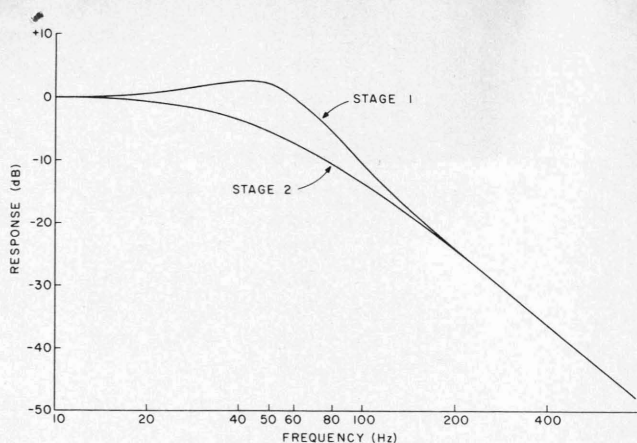


Fig. 9. Response of two filter stages.

Stage 1: $f = 50$ Hz; desired gain $G = 2$; from Table 1 we need $\xi = .383$; I chose $C2 = 0.01$ μ F. From the equations, I got the following:

$R1 = 60,956$ Ohms
 $R2 = 121,912$ Ohms
 $R3 = 40,637$ Ohms
 $C1 = 0.2045$ μ F

Since I did not have exactly these values, I let $R1 = 56k$, $R2 = 120k$, $R3 = 39k$, and $C1 = 0.22$ μ F.

Stage 2: $f = 50$ Hz;

desired gain $G = 2$; from Table 1 we need $\xi = 0.924$; I chose $C2$ again at 0.01 μ F. From the equations, I got the following:

$R1 = 147,059$ Ohms
 $R2 = 294,118$ Ohms
 $R3 = 98,039$ Ohms
 $C1 = 0.0351$ μ F

The values actually used were $150k$ for $R1$, $270k$ for $R2$, $100k$ for $R3$, and 0.032 μ F for $C1$ (0.022 in parallel with 0.01). The complete circuit is shown in Fig. 8.

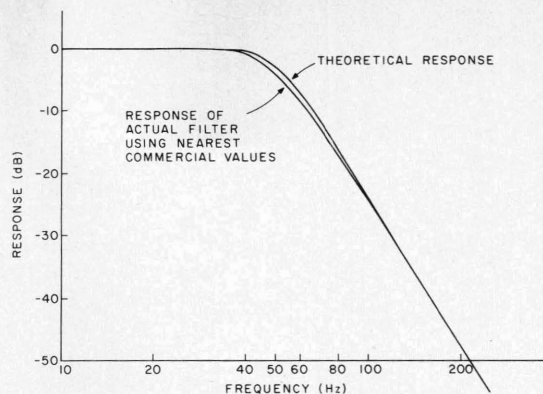


Fig. 10. Actual low-pass filter response.

The frequency response calculated for each of the two stages is shown in Fig. 9. Fig. 10 gives the theoretical response for the four-pole Butterworth filter, along with the frequency response actually measured on the circuit using the commercial equivalent values given in Fig. 8. Although the test equipment was not particularly fancy—the output above about 150 Hz was too low to be measured by my

meter—this shows that the equations seem to work. Only the gain was slightly off; theoretical gain should have been 4 (2 in each stage), while the total gain measured only a bit more than 3 . Nevertheless, the total gain could have been easily adjusted by changing the value of $R1$ in either stage. In general, the performance of these low-pass filters is close to what you would expect from the equations. ■

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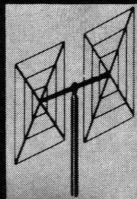
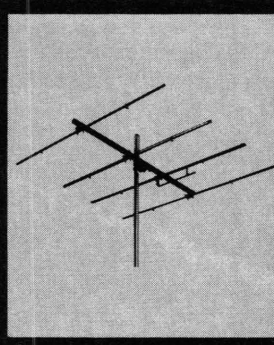
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Microstrip

—magical PC technique explained

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Microstrip is a transmission line, much like coax or twinlead, except that it is fabricated not from wire or tubing, but generally by etching traces on a printed circuit board. Now, it's a cinch the

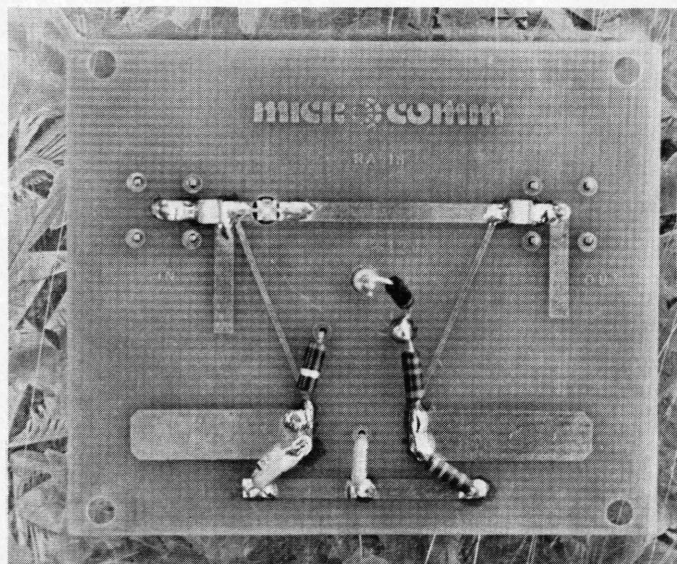
average radio amateur isn't about to run a long, thin strip of printed circuit board up his tower, so the actual application of microstrip transmission lines is likely to be something other than feeding antennas. In fact, microstriplines are especially useful in developing couplers, baluns, power dividers and combiners, matching transformers, capacitive and inductive reactances, and a whole host of structures which utilize the distributed properties of transmission lines in lieu of "lumped" components.

The recent acceptance of microstrip techniques by radio amateurs has allowed PC layouts to be developed such that the traces used to interconnect components will themselves be used as circuit components. I have recently presented microstripline circuits for amplifiers,^{1,2} mixers,^{3,4,5} filters,⁶ and complete converters.⁷ In each case, successful duplication of the designs re-

quired only selecting the proper printed circuit material and adhering closely to the published layout and dimensions. This article is in response to numerous inquiries from radio amateurs who had successfully duplicated some of my designs, but wanted to know how the magical PC trace dimensions were determined.

Transmission Line Parameters

Any transmission line, be it coax, twinlead, microstrip, or what-have-you, can be described in terms of its characteristic impedance (Z_0) and length. Characteristic impedance is the terminating impedance seen looking into an infinitely-long section of the transmission line and is a function of the dimensions of the conductors, their orientation with respect to one another, and the dielectric constant of any material separating them. Transmission line length may refer to a physical di-



Typical microstripline preamplifier. In this circuit, the PC traces perform as matching transformers, inductors, rf chokes, and bypass capacitors. This particular preamplifier was designed for use in receiving weather satellites at 1.7 GHz, although the author has built similar units to cover the 420 MHz, 1215 MHz, and 2300 MHz ham bands.

mension, but more often denotes the phase delay (measured in electrical degrees) of a signal being propagated down the line at a particular operating frequency. Such an electrical length is, of course, a function of the velocity at which the signal is propagating down the line; hence, if we specify a transmission line's velocity of propagation, we can generally calculate the electrical length of any physical line at any desired frequency. Since the velocity of propagation of electromagnetic radiation is free space equals the speed of light (a universal constant), and since this velocity cannot be exceeded by a signal propagated down an actual transmission line, it is common practice to specify for any transmission line its *relative velocity of propagation* (with respect to the speed of light), or *propagation constant*. From this decimal number, the electrical length at a given frequency is easily found for any physical length of transmission line.

I have recently published design techniques⁸ which determine the characteristic impedance and electrical length required of transmission lines to be inserted at specified locations within amplifier circuits. Similar techniques exist for other types of circuit designs. What remains is to determine the required dimensions and layout for achieving the desired transmission lines. Since microstriplines can be fabricated over a wide range of characteristic impedances and at practically any length (limited only by circuit board size restrictions), microstrip transmission lines would seem an ideal medium for most circuit design, from frequencies of a few hundred MHz up through

several tens of GHz.

Microstripline Variables

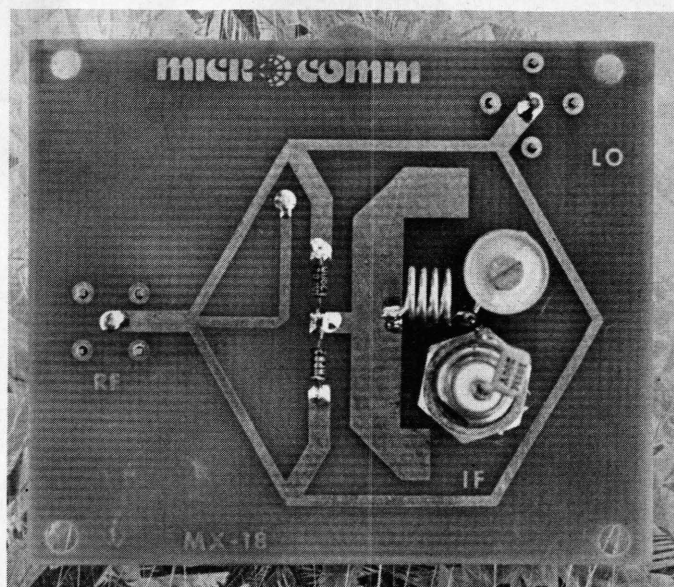
Any transmission line may be considered as a combination of series inductive and shunt capacitive elements, or sections, acting together as a delay line. The characteristic impedance of the resulting line can be shown to be proportional to the ratio of inductance to capacitance for a unit length.⁹ Algebraically,*

$$Z_0 = \sqrt{L/C}$$

Consider Fig. 1, which is a representation of a typical microstrip transmission line etched from double-clad printed circuit stock. The thin strip atop the board, with respect to the unetched ground plane, comprises the transmission line. A signal injected (or "launched") into the strip at one end will emerge, after a given propagation delay, from the other end. Since the above equation holds for at least a first-order approximation, determining the characteristic impedance of the microstrip transmission line in Fig. 1 involves merely quantifying the inductance and capacitance per unit length.

Like any conductor, the etched trace in Fig. 1 exhibits an inductance per unit length which is inversely proportional to its width, W , and thickness, t . The capacitance between the strip and the ground plane varies directly with the line width, W , varies directly with the dielectric

*Strictly speaking, this relationship only holds for transmission lines propagating signals in the transverse electromagnetic (TEM) mode—that is, with electric and magnetic lines of force both at right angles to each other and at right angles to the direction of propagation. Fortunately, microstriplines closely approximate TEM propagation in most applications.



Balanced mixer. Here microstriplines are used to introduce rf and LO signals to a pair of mixer diodes (center of board) in the proper phase relationship for linear heterodyne conversion to take place. The PC traces also furnish impedance matching i-f filtering and dc return for the mixer diodes. Again, this particular mixer is optimized for weather satellite reception at 1.7 GHz, but similar designs have been used for transmit or receive conversion in any of the amateur microwave bands.

constant, ϵ_r , of the material separating the plates of the capacitor, and varies inversely with the spacing, h , between the plates.

Thus a series of proportionalities can be developed:

$$Z_0 = \sqrt{L/C}$$

$$L \propto \frac{1}{Wt}$$

$$C \propto \frac{W\epsilon_r}{h}$$

Therefore,

$$Z_0 \propto \sqrt{\frac{1/Wt}{W\epsilon_r/h}}$$

where α (the Greek lower-case alpha) indicates that the quantities are proportional rather than equal.

Inverting the denominator within the radical sign and multiplying it by the numerator, we get:

$$Z_0 \propto \sqrt{\frac{h}{W^2\epsilon_r t}}$$

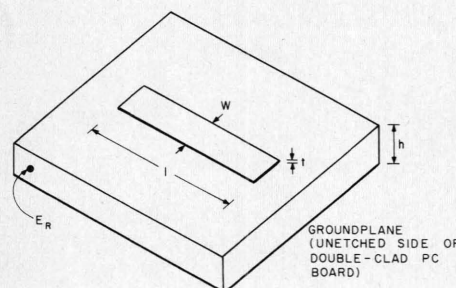


Fig. 1. Representation of a typical microstrip transmission line etched from double-sided printed circuit board. Variables affecting the line's electrical performance include the width, W , of the microstripline, the height, h , of the strip above the ground plane (this being related to board thickness), the thickness, t , of the strip metallization, the length, L , of the strip, and the relative dielectric constant, ϵ_r , of the substrate.

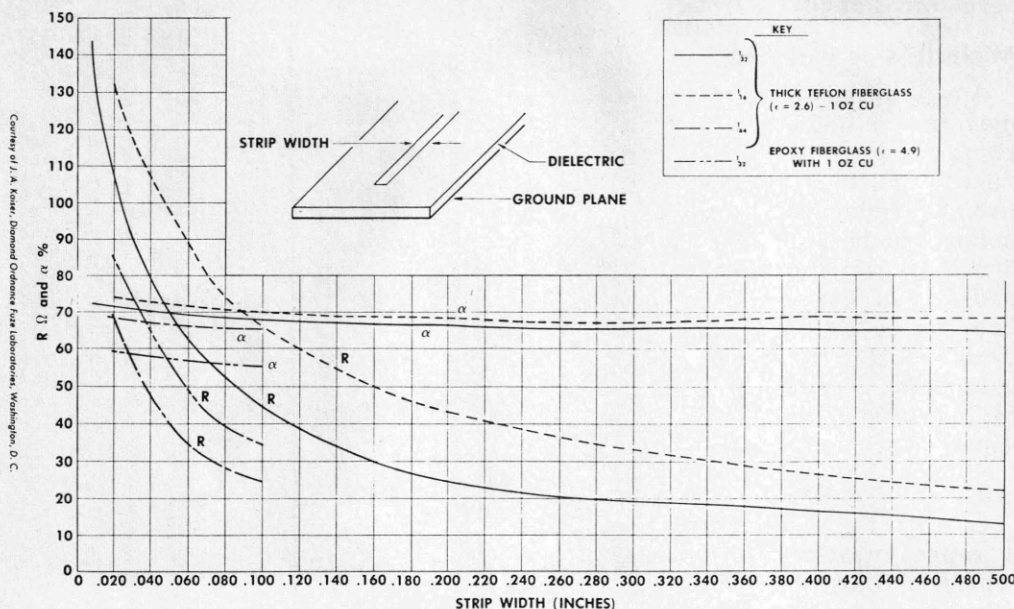


Fig. 2. Measured microstripline impedance and propagation constant versus strip width.

which at least allows us to draw some general conclusions about microstripline dimensions, if not determine them directly.

Generalization #1:

Wide lines have low characteristic impedance, while narrow lines exhibit high characteristic impedance.

Generalization #2:

For a given characteristic impedance, the required line width is narrower for thin substrates

than it is for thick substrates.

Generalization #3:

For a given characteristic impedance, line width will be greater on low dielectric-constant materials than it will with higher dielectric constants.

Generalization #4:

Increasing the thickness of the metallization tends to decrease the characteristic impedance of the line (this will be discussed in greater detail in a later section of this article).

A similar series of proportionalities can be developed for describing the velocity of propagation along a microstrip transmission line, starting from the relationship:⁹

$$v = \frac{1}{\sqrt{LC}}$$

Without belaboring the calculations, let us generalize:

Generalization #5:

Velocity of propagation decreases with increasing dielectric constant.

Generalization #6:

Velocity of propagation increases slightly as microstriplines become more narrow (that is, for higher characteristic impedances).

Generalization #7:

For a given dielectric constant, as long as strip width is adjusted to hold characteristic impedance constant, propagation velocity is independent of substrate thickness.

Since the computations involved in actually determining dimensions for a particular microstripline are quite involved and often fraught with error, the above approximations not only lend themselves to a better understanding of the mechanism of microstrip, but they also provide a "test of reasonableness" for evaluating the results of a more rigorous computational analysis.

Quantifying Microstripline Behavior

A set of equations frequently encountered in the engineering literature¹⁰ which provides a good first-order approximation of the characteristic impedance and propagation velocity of microstriplines is as follows:

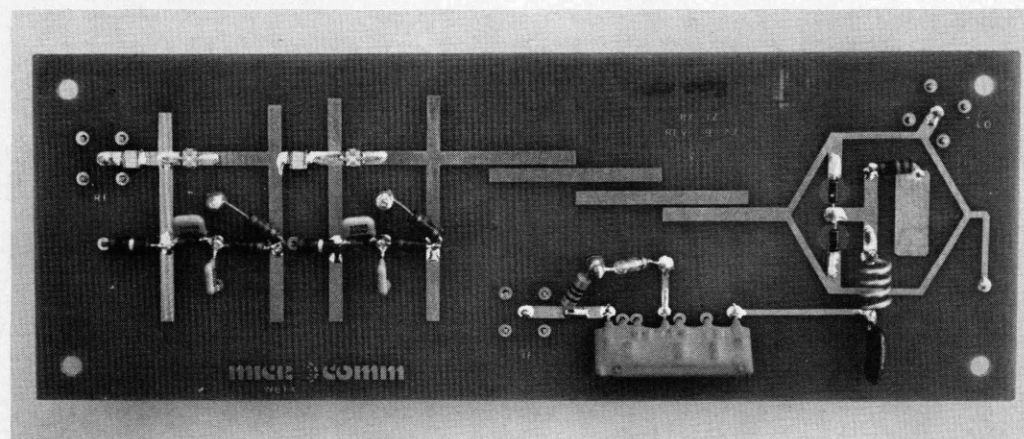
$$Z_0 = \frac{377(h/W)}{\sqrt{\epsilon_r}}$$

and

$$v = \frac{c}{\sqrt{\epsilon_r}}$$

where W , h , and ϵ_r are as defined in Fig. 1, c represents the velocity of propagation of radiant energy in free space, and the number 377 is an expression for the characteristic impedance of free space in Ohms.*

Although the above for-



Integrated microstrip assembly. This board is a complete front end for the reception of TV signals from satellites at 2.6 GHz. Microstriplines provide matching and bias feeds for the two preamplifiers (left) and i-f amplifier (bottom right), furnish ample bandpass filtering for spurious-free reception (center), and phase rf and LO signals in a balanced mixer (far right). This particular board provides 30 dB of conversion gain, a 3 dB noise figure, and better than 20 dB of image frequency rejection. A similar unit has been used for successful ATV reception in the 2.4 GHz band.

* $Z = \sqrt{\mu_0/\epsilon_0}$, from reference 9. Since μ_0 , the permeability of free space, is approximately $4\pi \times 10^{-7}$ Henrys per meter, and ϵ_0 , the permittivity of free space, is approximately $1/36\pi \times 10^{-9}$ Farads per meter, it can be shown that $Z_0 \approx 377\Omega$.

formulas yield results which conform to the generalizations introduced previously, they are at variance with measured data on actual microstriplines. This is because the ideal formulas assume true TEM propagation, with the conductors completely immersed in the dielectric medium, and ignore such ever-present anomalies as fringing capacitance (capacitive coupling between the microstrip and the outside world) and flux leakage (mutual inductive coupling between the strip and its surroundings). Throughout the 1950s and '60s, numerous studies were performed by Cohn, Wheeler, Sobol, Schneider, and others to more completely characterize the behavior of microstriplines under "real-world" conditions. Documentation of these investigations may be found in references 11 through 22. The calculations are involved but have enabled a number of graphical aids for the dimensioning of microstriplines to be developed.

Graphical Analysis—Wheeler's Charts

The technique most commonly used by microwave engineers to dimension microstriplines involves a set of graphs known as "Wheeler's Charts," after Harold A. Wheeler, the author of references 15 and 16. The charts, published in numerous technical journals, were not actually developed by Wheeler, but are generally based upon a set of equations derived by him. Wheeler's Charts take several forms, and may be used for determining the proper width for a microstripline of a desired characteristic impedance, for determining the resulting velocity of propagation, or both. Like any graphical design tool,

Wheeler's Charts are only approximate, being limited in resolution by their finite size. Nonetheless, data derived from these charts has provided a good starting point for literally thousands of successful microwave designs.

Fig. 2 is one form of Wheeler's Chart, published for several years in the *Microwave Engineer's Handbook*.²³ From it, one can determine to a high degree of accuracy the required strip width for a desired characteristic impedance and the resulting propagation velocity (as a percentage of the speed of light) for two particular types of substrate material in three popular thicknesses. No information as to effects of trace thickness is given in this chart, but it seems to hold for 1 oz. (0.0014" thick) and 2 oz. (0.0028" thick) copperclad PC laminate.

The Wheeler's Charts in Figs. 3, 4, and 5 are more general in that they allow microstriplines to be dimensioned independently of substrate thickness or dielectric material. These charts were published a couple of years ago by Communications Transistor Corporation²⁴ and formed the basis of most of the microstrip dimensions presented in my various construction articles. To use Fig. 3, it is necessary to know the relative dielectric constant (ϵ_r) of the substrate material (its permittivity relative to that of free space). For a given characteristic impedance, the chart gives the ratio of required microstripline width to height (the dielectric thickness). Thus the required strip width is found merely by multiplying this ratio by the actual substrate thickness used, making this chart applicable to any desired substrate thickness.

Figs. 4 and 5 similarly

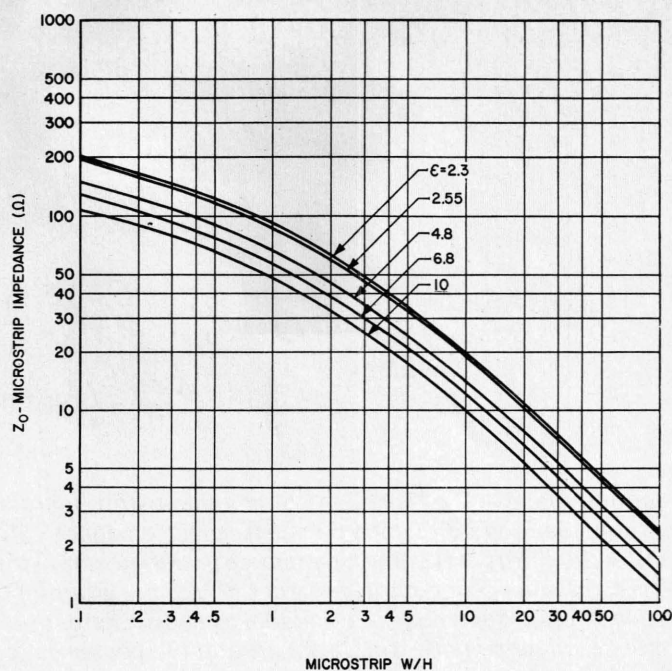


Fig. 3. Microstrip impedance versus width/height.

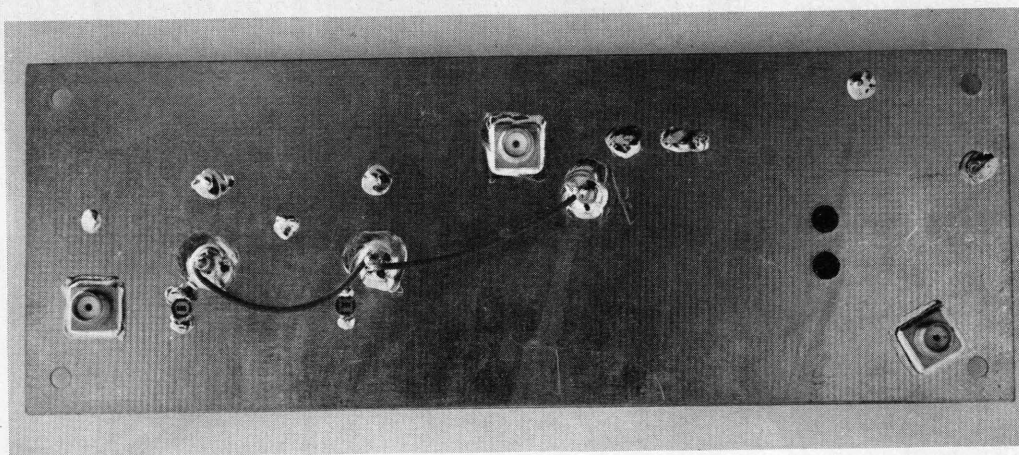
employ the width-to-height ratio of a microstrip and are used indirectly to determine propagation velocity. To understand their use, it is necessary to first introduce the parameter which the charts erroneously call relative dielectric constant, but which more properly should be referred to as effective dielectric constant, ϵ_{eff} .

Effective Dielectric Constant

The first-order approximations given earlier for microstripline characteristic impedance and propagation velocity would hold if the dielectric material completely surrounded the strip conductor and completely contained all magnetic and electrical lines of force. In fact, the dielectric constant of the substrate (between the strip and the ground plane) and that of the material above the strip (typically air) are sufficiently different that a sizable dielectric discontinuity exists. This is further aggravated by the fringing capacitance and magnetic flux leakage problems mentioned earlier. Now, if

all these discontinuities were taken into account, a correction factor could be developed, tacked onto the textbook formulas for Z_0 and v_{rel} , and all would be well. Unfortunately, this correction factor to Z_0 is actually a function of Z_0 , which is why the formulas developed by Wheeler, Sobol, Schneider, and others are so complex. But assuming the actual Z_0 of a stripline were found, and, if it were compared to the assumed Z_0 from the textbook equation, a correction factor could be derived for each particular microstripline measured. That correction factor could then be multiplied by ϵ_r , the permittivity of the dielectric relative to that of free space, resulting in the parameter which I call ϵ_{eff} , effective dielectric constant.

Since, for a given microstripline dimension, ϵ_{eff} encompasses all the corrections necessary for fringing capacitance, flux leakage, and dielectric discontinuity, the first-order formulas could be used if ϵ_{eff} were substituted for ϵ_r throughout. The only problem remaining is to deter-



Ground plane—The “flip side” of any microstrip board remains unetched, to serve as a ground plane against which transverse electromagnetic (TEM) waves can propagate down the etched strip. It is common practice to design microstrip boards so that dc power feed-throughs and rf connectors project through the ground-plane side. When mounting coax connectors through a ground plane, as shown here, don’t forget to remove a bit of the ground-plane metallization from around the center pin or you’ll end up shorting out all signals!

mine the value of ϵ_{eff} , which leads us to another generalization.

Generalization #8:

Since all flux lines surrounding the stripline pass both through the substrate (of dielectric constant ϵ_r) and also through air (dielectric constant $<\epsilon_r$), the effective dielectric constant, ϵ_{eff} , will always be less than ϵ_r .

Figs. 4 and 5 list, for various substrates of dif-

ferent relative dielectric constant, ϵ_r , the effective dielectric constant, ϵ_{eff} , as a function of width-to-height ratio (which was itself a function of characteristic impedance). From ϵ_{eff} , it is relatively easy to determine v_r , relative propagation velocity as compared to the speed of light, from the formula:

$$v_r = \frac{1}{\sqrt{\epsilon_{eff}}}$$

Effects of Metallization Thickness

Nowhere in Wheeler’s Charts is there any correction for t , the thickness of the strip conductor and ground-plane metallization. Actually, generalization #4 notwithstanding, varying the metallization thickness over a fairly wide range has virtually no effect upon the properties of the microstrip. This is because skin effect forces

most of the current to flow at or near the surface of a conductor, and this effect is amplified as frequency increases. As long as the metallization thickness is sufficient to support the required current flow (and, at microwave frequencies, a few microns will suffice), circuit operation is relatively independent of trace thickness.

Yet it is important to know the metallization thickness when dimensioning microstriplines. This is so because the specified thickness of copperclad printed circuit laminates is generally inclusive of metallization thickness, but the microstripline formulas or charts require a knowledge of the substrate thickness alone.

Consider a printed circuit board 1/32” (0.79 mm) thick, double-clad with 2-ounce copper. The copper thickness will be 0.0028” (0.071 mm) on each side, leaving only 0.65 mm of dielectric thickness. Obviously, any microstripline calculation which ignores metallization thickness, and assumes a 0.79 mm dielectric, will be in error.

As mentioned previously, Fig. 2 seems to provide adequate results with either 1- or 2-ounce copper thickness. Nonetheless, when utilizing the width-to-height ratios of Figs. 3, 4, or 5, or any time high precision is required, the actual metallization thickness should be taken into account when determining the height, h , of a microstrip above the ground plane.

Characteristic Impedance Limitations

There is a finite range of widths (hence characteristic impedances) over which an etched microstrip will perform as a TEM transmission line. As a line becomes very wide, its capacitance per unit

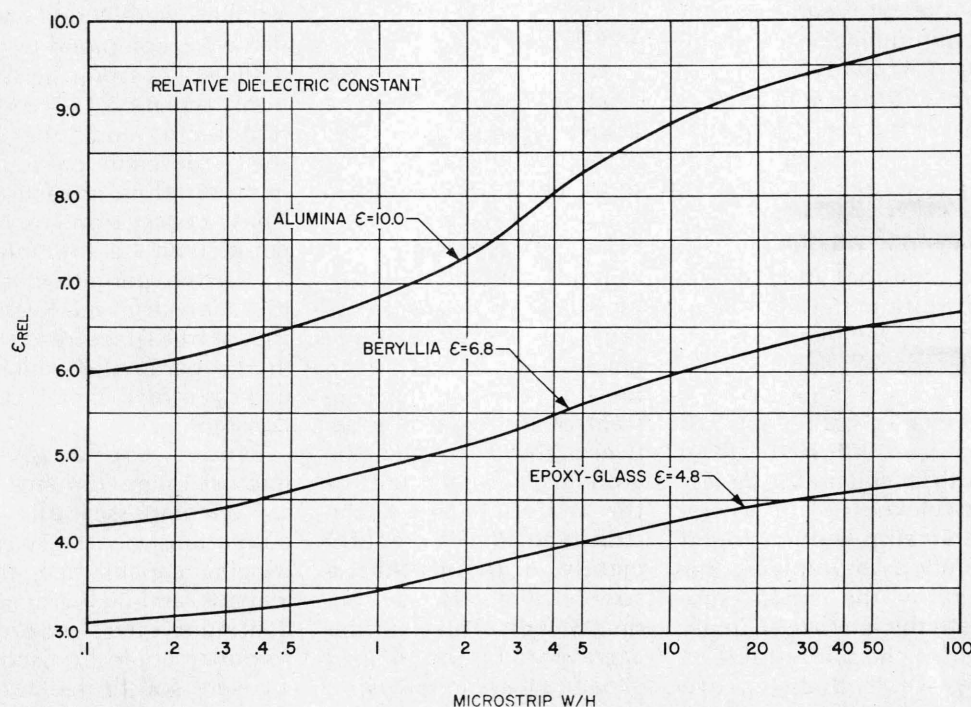


Fig. 4.

length becomes so great as to completely dominate its behavior. The inductance per unit length becoming insignificant, the line can no longer behave as a transmission line, but rather forms a lumped shunting capacitance to ground. This transition begins to occur when the width of the strip exceeds about 10 times its height above the ground plane. Similarly, for very narrow lines, the capacitance per unit length becomes insignificant, the inductance per unit length predominates, and the line no longer supports TEM propagation but instead degenerates to a series inductor. This transition seems to occur when the height of the strip above ground exceeds the strip's width by about a factor of ten.

From Fig. 3, it can be seen that the above limitations tend to restrict the useful range of microstrip-line characteristic impedances to about 15 to 150 Ohms. Needless to say, the ability to construct TEM transmission lines of any desired characteristic impedance between these two extremes is hardly a severe limitation and will likely accommodate just about any design requirement.

Length Limitations

Microstriplines which are very short relative to their width similarly tend to act as shunt capacitances rather than transmission lines. This occurs because the signal, rather than propagating along the length of the line, actually propagates from the center of the line out toward the edges. It is apparent that with signals traveling in this direction, they are no longer perpendicular to the magnetic and electric fields, so the strip no longer acts as a TEM transmission line.

As a rule, no microstrip-line should be designed into a circuit such that its width exceeds its length. In circuits requiring a short electrical length of transmission line, a certain amount of ingenuity is required. In many applications, it is possible to take advantage of the property of repeating impedances at half-wave intervals. For example, if a circuit should require a shunt 0.1-wave-length transmission line, but if the frequency were sufficiently high to make the microstripline short relative to its width, a 0.6-wave-length transmission line may well suffice.

In applications where it is not practical to add an integral half wavelength to the transmission line, it may be possible to design with a longer transmission line of some other impedance or perhaps a short length of transmission line with a higher characteristic impedance (hence narrower strip width). Where such circuit modifications are not possible, it may be necessary to redesign the circuit onto a thinner substrate. This has the effect of reducing the width of all microstriplines in the circuit.

Since attenuation in a microstripline, as in any transmission medium, is a direct function of transmission line length, it should be obvious in designing circuits that no line should be made longer than absolutely necessary. An exception is the use of long, lossy sections of transmission line as attenuators in a circuit.

Selecting Stripline Laminates

Throughout the microwave industry, the use of such prosaic materials as fiberglass-epoxy printed circuit board as microstripline laminates is met with scorn. Nonetheless, I have

Z_0	1/32" TFE, 1 oz Cu			1/16" TFE, 1 oz Cu		
	W	ϵ_{eff}	v/v_0	W	ϵ_{eff}	v/v_0
15	.377	2.35	.65	.791	2.34	.65
20	.268	2.29	.66	.564	2.29	.66
25	.204	2.25	.67	.429	2.25	.67
30	.162	2.21	.67	.340	2.21	.67
35	.132	2.18	.68	.277	2.18	.68
40	.109	2.16	.68	.231	2.15	.68
45	.092	2.13	.68	.195	2.13	.68
50	.079	2.11	.69	.166	2.11	.69
60	.059	2.07	.70	.124	2.07	.70
70	.045	2.03	.70	.095	2.03	.70
80	.035	2.00	.71	.074	2.00	.71
90	.028	1.97	.71	.059	1.98	.71
100	.022	1.95	.72	.047	1.95	.72
110	.018	1.92	.72	.038	1.92	.72
120	.014	1.91	.72	.030	1.91	.72
130	.011	1.90	.72	.024	1.91	.72
140	.009	1.90	.73	.019	1.90	.73
150	.007	1.89	.73	.015	1.89	.73

Table 1: Width (in decimal inches), effective dielectric constant, and velocity factor for various characteristic impedance microstriplines etched on two different thicknesses of fluorocarbon-dielectric ($\epsilon_r = 2.55$) printed circuit stock, double-clad with 1 oz. copper. This table is also applicable to various low-dielectric-constant polyolefin and glass-derived printed circuit materials, as discussed in the text.

achieved a reasonable degree of success with microstriplines on such inexpensive circuit stock, as reported in my previous articles.

The attitude of the industry is not without some basis, as process controls in the manufacture of garden-variety glass-epoxy PC board are somewhat lacking. The laminate thickness and dielectric constant will vary widely between vendors and for different production runs of board from the same manufacturer, as reported by Motorola.²⁵ Nonetheless, it is possible to design the circuit to be forgiving of such anomalies, especially if variable tuning elements are incorporated in the circuit for impedance matching. Further, Sobol's microstrip equations (reference 18) permit strips to be dimensioned for any board thickness and dielectric constant encountered, as long as the material to be used is measured.

Whenever I purchase a quantity of glass-epoxy circuit board, I make it a

point to strip back the metallization on both sides of a sample and measure the actual dielectric thickness. I have never observed deviations from optimum of more than a few percent, so the manufacturer's published thickness is probably close enough for dimensioning striplines. As for relative dielectric constant, ϵ_r , I measure it by determining the capacitance of a scrap of double-sided laminate whose dimensions are precisely known. This is most easily done on an automatic RCL bridge, but any available capacitance-checker should suffice. Dielectric constant is then found from the formula:

$$\epsilon_r = \frac{hC}{8.85A}$$

where C is capacitance in pF, h is dielectric thickness in meters, and A is the area of either plate in square meters.

If the resulting dielectric constant is reasonably close to the theoretical 4.8 for fiberglass-epoxy PC board, go ahead and design

1/32" glass-epoxy 1 oz. Cu				1/16" glass-epoxy 1 oz. Cu			
Z ₀	W	ϵ_{eff}	v/v_0	W	ϵ_{eff}	v/v_0	
15	.262	4.15	.49	.551	4.15	.49	
20	.184	4.01	.50	.387	4.01	.50	
25	.138	3.91	.51	.290	3.91	.51	
30	.107	3.82	.51	.226	3.82	.51	
35	.086	3.74	.52	.181	3.74	.52	
40	.070	3.68	.52	.148	3.68	.52	
45	.058	3.61	.53	.123	3.62	.53	
50	.048	3.55	.53	.103	3.56	.53	
60	.034	3.45	.54	.074	3.46	.54	
70	.026	3.37	.54	.055	3.37	.54	
80	.019	3.28	.55	.041	3.29	.55	
90	.014	3.23	.56	.031	3.24	.56	
100	.011	3.22	.56	.023	3.22	.56	
110	.008	3.20	.56	.017	3.20	.56	
120	.006	3.18	.56	.013	3.18	.56	
130	.004	3.15	.56	.010	3.16	.56	
140	.003	3.13	.56	.007	3.14	.56	
150	.002	3.11	.57	.005	3.12	.57	

Table 2. Width (in decimal inches), effective dielectric constant, and velocity factor for various characteristic impedance microstriplines, etched on two different thicknesses of fiberglass-epoxy dielectric ($\epsilon_r = 4.8$) printed circuit stock, double-clad with 1 oz. copper.

microstriplines from Wheeler's Charts or from the tables included in this article. If not, you have the option of selecting another batch of PC stock and trying again or designing your strip dimensions around the actual ϵ_r of the material on hand.

The majority of commercial microstripline users employ a type of circuit

board exhibiting tightly controlled dielectric properties. Traditional materials include fluorocarbon laminates (such as Teflon, a DuPont trade name for tetra-fluoro-ethylene), polyolefins (such as polyethylene), or glass derivatives such as "duroid" or "rexolite." All of these materials exhibit a much lower dielectric constant

than fiberglass-epoxy (on the order of 2.5, as opposed to 4.8), which results in wider, longer striplines—a decided advantage at higher frequencies where strips might otherwise become so short as to be unmanageable. Although these stripline laminates are quite a bit more costly than glass-epoxy, they offer exceptionally consistent properties and excellent performance well beyond ten GHz. Glass-epoxy, on the other hand, is only marginally useful at 2300 MHz and becomes extremely lossy beyond 3 GHz.

Military and aerospace microwave circuitry is frequently fabricated by depositing gold traces on a ceramic substrate of highly controlled dimensions and dielectric properties. The most popular of these substrates is alumina, which has an extremely high dielectric constant (around 10). The very high ϵ_r considerably reduces microstripline dimensions, which is a definite asset in high-density applications such as microwave integrated circuits, although it greatly increases the dimensional precision required both in design and fabrication. Until recently, such high- ϵ_r materials were completely beyond the reach of the average experimenter. A new microstripline laminate from 3M Company, called ϵ silam 10, promises to change that. This board has a ceramic-impregnated dielectric material whose dielectric properties match those of alumina, but which can be machined like conventional printed circuit material. The board is supplied double-clad with 1-ounce copper and can be etched with either ferric chloride or chromic-sulphuric acid. Although the cost is quite high (on the order of \$1 per square inch), ϵ silam 10 promises

to make high-density techniques available to the interested microwave amateur without requiring investment in exotic processing equipment.

Introducing Sobol's Tables

Wheeler's Charts, as seen in Figs. 2 through 5, provide a convenient technique for determining microstripline dimensions for a desired characteristic impedance and electrical length. However, it is frequently more convenient to have this information available in tabular form, especially when a limited number of standard dielectric types and thicknesses are used. I recently developed a set of such tables for finding width, effective dielectric constant, and velocity factor of microstriplines over a wide range of characteristic impedances for six different frequently-encountered laminates.

Since the calculations are quite involved, require multiple iterations and conditional branching, and include parameters which are interactive, it was decided to employ a programmable pocket calculator (in this case, the Hewlett-Packard Model 25). From a wide field of available equation sets, I selected Sobol's equations from reference 18, primarily because they lent themselves to entry within the limitations of my calculator's 49-step program capacity.* The calculator programs are available to anyone interested in such

*It is recognized that Sobol's equations, having been derived more than a decade ago, are less precise than others published more recently. However, since the errors in microstripline dimensions utilizing Sobol's equations seldom exceed a few percent, they are considered entirely satisfactory for amateur (if not government) work.

.025" alumina deposited gold traces				.050" alumina deposited gold traces			
Z ₀	W	ϵ_{eff}	v/v_0	W	ϵ_{eff}	v/v_0	
15	.146	8.28	.35	.293	8.29	.35	
20	.100	7.95	.35	.201	7.96	.35	
25	.073	7.69	.36	.146	7.69	.36	
30	.055	7.46	.37	.111	7.46	.37	
35	.043	7.26	.37	.086	7.26	.37	
40	.033	7.06	.38	.068	7.09	.38	
45	.026	6.89	.38	.054	6.91	.38	
50	.022	6.77	.38	.045	6.78	.38	
60	.015	6.50	.39	.030	6.50	.39	
70	.010	6.43	.39	.020	6.43	.39	
80	.006	6.35	.40	.013	6.36	.40	
90	.004	6.28	.40	.009	6.30	.40	
100	.003	6.24	.40	.006	6.24	.40	
110	.002	6.18	.40	.004	6.18	.40	
120	.001	6.07	.41	.002	6.07	.41	
130				.001	5.97	.41	

Table 3. Width (in decimal inches), effective dielectric constant, and velocity factor for various characteristic impedance microstriplines of gold traces deposited on two different thicknesses of alumina ($\epsilon_r = 10.3$) substrate. This table is also applicable to the new 3M brand " ϵ silam 10" high-dielectric-constant printed circuit stock, as discussed in the text.

calculations,* but the results are presented in Tables 1, 2, and 3.

I had actually intended to name these tables after myself. After all, a great deal of time and effort went into writing the programs and computing the data. Then I had the unexpected pleasure of meeting Dr. Harold Sobol for the first time, at a recent International Microwave Symposium. I found Hal to be stimulating, personable, enjoyable—a "gentleman and a scholar" in the true sense of the expression. Thus I decided that the tables which I am presenting here, like Wheeler's Charts, should be named after the person who derived the formulas, rather than the person who applied them in a convenient form.

Conclusion

Microstrip transmission lines lend themselves to the design of impedance matching networks, coupling structures, and reactive circuit elements. Their usefulness extends from the VHF region far into the microwave spectrum. This article has presented several generalizations about the dimensioning of microstriplines, which may be verified by examining Sobol's Tables (Tables 1 through 3).

Actual dimensions for microstrip transmission lines of a desired characteristic impedance and electrical length can be determined graphically, from tables, or with the aid of a programmable calculator. Try microstripline for

*The HP-25 programs used to develop Tables 1 through 3 form a part of a microwave design program library developed by the author and are available for a nominal charge. For details, send a stamped self-addressed envelope to Microcomm, 14908 Sandy Lane, San Jose CA 95124.

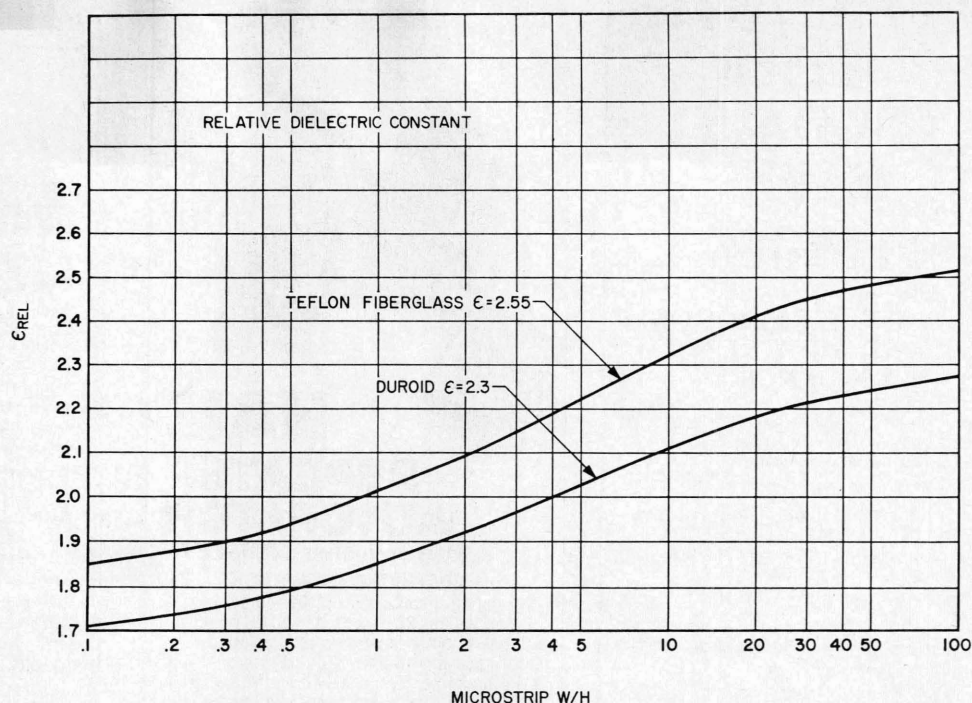


Fig. 5.

your next microwave design or construction project. It's simpler than you realized and more rewarding than you can imagine! ■

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